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January 16, 1963

RADIO SET AN/TRC-56
Final
Technical Report

Quentin Veit
George Mackiw
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PHILCO CORPORATION
A Subsidiary of Ford Motor Company
Communications and Electronics Division
4700 Wissahickon Ave.
Philadelphia 44, Pennsylvania

Philco No. 2304
Contract AF 30(602)2331

Prepared
for
Rome Air Development Center
Research and Technology Division
Air Force Systems Command
United States Air Force

Griffiss Air Force Base
New York

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ABSTRACT

This report describes experimental models of Radio Set AN/TRC-56, developed for Rome Air Development Center by Philco Corporation under Air Force Contract AF 30(602)-2331. The AN/TPC-56 Radio Set consists of a radio unit and a multiplex unit.

The portable radio unit operates in either the 7.125 - 8.4 gc or 14.0 - 15.4 gc frequency bands. It is tripod-mounted and has a self-contained antenna. It is completely solid-state except for two klystrons which serve as the transmitter and local-oscillator tubes. The radio unit is capable of operating with various types of multiplex in addition to its companion AN/TRC-56 multiplex. It is also capable of carrying television signals.

The AN/TRC-56 multiplex unit is a frequency-division multiplex equipped with twelve 4-kc voice channels and two 20-kc data channels, with expansion capability for an additional 10 data channels. Each voice channel is equipped with a termination unit providing two-wire, four-wire, a-c ringdown, and E & M dialing options. The unit is completely solid-state.

This report contains the physical and functional descriptions, design goals, and test results for the experimental models. In addition, recommendations for future equipments are made.

PUBLICATION REVIEW

This report has been reviewed and is approved.

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SECTION I

INTRODUCTION

The AN/TRC 56 Program consisted of the development of a line-of-sight microwave equipment with a multiplex terminal unit. The radio unit was to be mounted on a tripod assembly with the antenna an integral mechanical part of the radio package. The frequency of transmission was to take place in two different bands; 7.125 gc-8.4 gc, and 14.0 gc-15.4 gc. Both units, the radio and multiplex, were to be unattended, portable, solid-state equipments which could be subjected to extreme environments.

The purpose of this report is to discuss the achievements and shortcomings of the developed equipment, its physical and functional aspects, and the circuit methods employed. Sections II and III are discussions of the multiplex and radio, and contain detailed information of each equipment. Section IV is a discussion of the units as a system. The bulk of information on both units was separated into two sections since, aside from level requirements at baseband frequencies, the specifications are identified on a per-unit basis.

SECTION II

MULTIPLEX

A. GENERAL

1. Mechanical Description

The AN/TRC-56 MULTIPLEX is an entirely solid-state equipment designed for field use under extreme environmental conditions. The multiplex is contained in one integral ruggedized package. Most of the circuitry is placed on printed boards and has been designed for quick access. Shock frames provided with the unit are used when the equipment is in transit. Setup of the equipment requires removal of the front cover and the plate assemblies. After the equipment has been set up, the front cover can be replaced to environmentally seal the unit. Internal forced-air convection is used to cool the unit to ensure longer life for the active components. Interconnection of the various assemblies is made at the back of the unit. There are four major groups of subassemblies: the Power Supply, the Carrier Generator Assembly, the Signaling and Termination Assemblies, and the Transmit-Receive Assemblies.

The Power Supply Assembly is in the center portion of the equipment and is entirely removable as an integral unit from the front of the unit. The Power Supply Assembly consists of the Metering Facilities, the Transmit Baseband Amplifier, and the Power Supply. The Carrier Generator Assembly is located at the center-back of the unit and is removable from the back of the unit. The Signaling and Termination Assemblies, which are printed boards, slide on tracks for easy removal and are placed side by side at the bottom of the unit. The Transmit-Receive Assemblies are arranged in two rows above the Signaling and Termination Assemblies and are also made for quick removal. The row of 12 Transmit-Receive Assemblies immediately above the Signaling and Termination Assemblies are the Voice Channels. Each Transmit-Receive Assembly of the Voice Channel is associated with a Signaling and Termination Assembly directly underneath. The top row of Transmit-Receive Assemblies are the data channels. Both the Signaling and Termination, and the Transmit-Receive Assemblies are removable from the front of the unit after the front cover and plate assemblies have been taken off. See Figure 1 for the physical location of these assemblies.

The Signaling and Termination Assemblies are all identical and can be interchanged. Different options provided by the Signaling and Termination assemblies are chosen by removing the card and operating the switches to their proper positions. Adjustments of input and output levels on the Transmit and Receive cards are made with the printed board connected in the unit. These adjustments are accessible when the plate assemblies are removed. The plate assemblies hold the printed boards firmly in place.

The circuitry on the cards consists of printed copper with gold flash to ensure good solder connections. The boards are fungus proofed. Transistors are mounted in clips for mechanical rigidity. All tantalum capacitors are held firmly in place with an epoxy.

2. Functional Description

Figure 2 shows in simplified block form the interconnection of the assemblies in the Multiplex Equipment. The Transmit-Receive Assemblies are basically identical except for the frequencies at which filtering takes place. A Voice Channel without the Signaling and Termination Assembly is the same as a Data Channel except for its narrower bandwidth. The Signaling and Termination unit adapts a particular channel for E & M Dialing or AC Ringdown operation; and for Two-Wire or Four-Wire operation. The various options are chosen by positioning switches located on the assembly. For both AC Ringdown and E & M Dialing operation, the Signaling Termination Assembly produces a square-wave output which keys the transmitted channel carrier on and off at the Transmit-Receive Assembly. The Signaling and Termination Assembly receives the keyed carrier from the Transmit-Receive Assembly and, depending upon switch orientation, either turns a relay switch on and off for E & M operation or connects 20 cps ringing to the two-wire line for AC Ringdown operation. Audio signal in and out is connected to the Transmit-Receive Assemblies through the Signaling and Termination Assemblies which provide a 4-wire or 2-wire input.

On each Transmit-Receive Assembly, the input audio, or data, amplitude modulates a 150-kc carrier which is then filtered by a bandpass filter. The modulated signal is then heterodyned to its proper channel frequency. After filtering, the combined channel signals are amplified in the baseband amplifier. The combined baseband signal can now be transmitted over a radio link. The receiver portion of the Transmit-Receive assembly receives the baseband signal from a radio and after filtering the baseband signal to retrieve only the channel information, heterodynes this signal to 150 kc. The signal then is envelope detected and low-pass filtered. The low-pass filter is flat to 4 kc for Voice Channels and flat to 20 kc for data channels. After

additional amplification, the Voice Channel signals are fed to the Signaling and Termination Assembly from which they are fed out of the equipment on the 2-wire line or one pair of the 4-wire lines. The outputs of the Data Transmit-Receive Assemblies are fed directly out of the equipment.

The Carrier Generator Assembly provides the Transmit-Receive Assembly with the carriers necessary for modulation. Each Transmit-Receive Assembly is provided with 150 kc and a carrier which, when mixed with 150 kc, produces the channel carrier. Table I shows the frequencies which the Carrier Generator provides to the different channels.

The Power Supply Assembly consists of the Baseband Amplifiers, the Metering and Test Facility, and the Power Supply. The Metering and Test Facility is designed for field use so that the equipment can be set up to common operating levels without the use of external equipment. The Power Supply powers the blower and all the circuitry contained in the equipment.

3. Design Goals and Test Results

The design goals for the AN/TRC-56 are based on the requirements of RADC 2562A, Amendment 7. Table 2 presents in short form the tabulated specifications and test results. The subsequent tables (III through IX) and illustrations (Figures 3 through 38) present the test results in detail.

B. TRANSMIT-RECEIVE ASSEMBLIES

1. Mechanical Description

The Transmit-Receive Assemblies are 7 by 13-3/8 inch printed boards. The transmit portion of the assembly occupies approximately one third of the board space. An attempt was made to mechanically shield the the transmit side from the receive side by occupying the space between them with heavy grounds and encased units whose outer case was grounded. Circuits which operate at low frequencies are located the farthest from the connector. High-frequency circuits are located close to the connector. This was done to minimize radiation from board to board or within the board. Where possible,the encased units are also positioned near the edges of the card to give added strength. Two guides on the connector of the board promote smooth connection of the assembly. A Transmit-Receive patch board is provided for troubleshooting.

Table I Frequency Plan

Channel Number	Transmitter Reference Carrier (kc)	Transmitter Channel Carrier (kc)	Receiver Channel Carrier (kc)	Baseband Frequency (kc)
1	150	800	800	650
2	150	800	800	950
3	150	850	850	700
4	150	850	850	1000
5	150	900	900	750
6	150	900	900	1050
7	150	950	950	800
8	150	950	950	1100
9	150	1000	1000	850
10	150	1000	1000	1150
11	150	1050	1050	900
12	150	1050	1050	1200
13	150	200	200	50
14	150	200	200	350
15	150	250	250	100
16	150	250	250	400
17	150	300	300	150
18	150	300	300	450
19	150	350	350	200
20	150	350	350	500
21	150	400	400	250
22	150	400	400	550
23	150	450	450	300
24	150	450	450	600

Table II Specifications

Specification Item Number	Specification	Measured
3.5.3.1.1 (Data Chan.)	From 300 cps to 20 kc, the attenuation shall be within -1 db, +2 db referenced to 16 kc	This spec. was met by average frequency response of the data channels.
3.5.3.1.2 (Data Chan.)	Phase distortion coefficient from 600 cps to 19 kc shall not exceed 300 μsec	Less than 200 μsec
3.5.3.1.3	Single harmonic distortion within the band between 600 cps and 19 kc shall not be less than 30 db	Greater than 40 db at 1 kc; nominal value 50 db at 1 kc.
3.5.3.1.6	Crosstalk shall not be less than 60 db	Worst case, 51 db; nominal crosstalk, 67 db
3.5.3.1.7	Input impedance and input levels	600 ohms ± 10% (Z_{in})
3.5.3.2.10	a. Z_{in} 600 ohms b. Levels -10 dbm to +7 dbm	-8 to +7 dbm, 2-Wire Voice Chan. -25 to -10 dbm, Data Channel -25 to -10 dbm, 4-Wire Voice Channel
3.5.3.1.8	Output impedance and output levels	600 ohms ±10% (Z_{out})
3.5.3.2.11	a. Z_{out} 600 ohms b. Levels -10 dbm to +7 dbm	+7 dbm 2-wire and 4-wire Voice Chan.; +9 dbm Data Chan.
3.5.3.2.1 (Voice Chan.)	Frequency response from 300 cps to 3500 cps shall have attenuation variation of less than ±1 db about 1 kc	All channels within specification. Nominal frequency response was -0 db, +1 db about 1 kc

Table II Specifications (Continued)

Specification Item Number	Specification	Measured
3.5.3.2.2 (Voice Chan.)	Phase distortion coefficient, 165 μsec, 600 to 3200 cps Coefficient, 100 μsec 1 kc to 2.6 kc	130 μsec 600 to 3.2 kc 130 μsec 1 kc to 2.6 kc
3.5.3.2.2.1	Absolute delay, 300 cps to 3500 cps, 1.5 msec	Less than 1 msec
3.5.3.2.3	Single harmonic distortion from 300 cps to 3500 cps greater than 40 db, with 50 db as design goal	Greater than 40 db at 1 kc; nominal value of 50 db
3.5.3.2.6 (Voice Chan.)	Crosstalk shall not be less than 60 db	Worst case 51 db; nominal crosstalk 67 db
3.5.3.2.7	Echo attenuation shall be 25 db or more	Greater than 40 db when set for 0 dbm in, -6 dbm out, at 1 kc
3.5.3.2.12	Noise 32 dba at OLTP maximum	Worst case, 37.5 dba; Nominal, 26 dba

Table III AN/TRC-56 Multiplex Input Impedance

REFERENCE:

Philco Record Book: 11729
Sheet: No. 40
Date: August 14, 1962

TEST CONDITIONS:

Input level of voice channels (1 through 12) set at 0 dbm.
Input level of data channels (22 and 24) set at -16 dbm.
Voice channels connected for 2-wire operation.
Test frequency 1 kc.

TEST RESULTS:

MULTIPLEX A		MULTIPLEX B	
Channel Number	Z_{in} (ohms)	Channel Number	Z_{in} (ohms)
1	615	1	645
2	561	2	606
3	600	3	645
4	595	4	645
5	595	5	645
6	588	6	625
7	583	7	645
8	583	8	639
9	588	9	632
10	600	10	625
11	583	11	625
12	600	12	625
22	600	22	625
24	606	24	619

Table IV AN/TRC-56 Multiplex Noise Measurements

REFERENCE:

Philco Record Book: 11729
Sheet: No. 54
Date: August 22, 1962

TEST CONDITIONS:

Multiplex back-to-back.
Two-wire: 0 dbm in, -3 dbm out at 1 kc; channels 1 through 12.
Four-wire: -16 dbm in, -3 dbm out at 1 kc; channels 22 and 24.

TEST RESULTS:

	UNIT A TO UNIT B				UNIT B TO UNIT A			
	Noise		S/N Ratio		Noise		S/N Ratio	
Channel Number	F1A (dba*)	Flat (dbm)	F1A (db)	Flat (db)	F1A (dba*)	Flat (dbm)	F1A (db)	Flat (db)
1	22.0	-46.5	63.0	43.5	23.0	-44.5	62.0	41.5
2	24.8	-44.5	60.2	41.5	21.5	-49.0	63.5	46.0
3	25.4	-39.0	59.6	36.0	26.0	-52.0	59.0	49.0
4	24.2	-49.5	60.8	46.5	27.0	-44.0	58.0	41.0
5	23.2	-37.5	61.8	34.5	25.0	-35.5	60.0	32.5
6	27.5	-31.4	57.5	28.4	32.0	-31.0	53.0	28.0
7	24.0	-48.0	61.0	45.0	21.0	-53.5	64.0	50.5
8	31.6	-43.4	53.4	40.4	23.0	-50.5	62.0	47.5
9	25.8	-46.4	59.2	43.4	27.8	-43.4	57.2	40.4
10	27.6	-49.0	57.4	46.0	29.2	-47.0	55.8	44.0
11	31.8	-47.4	53.2	44.4	24.0	-45.5	61.0	42.5
12	37.5	-30.5	47.5	27.5	31.0	-42.5	54.0	39.5
22	23.2	-48.5	61.8	45.5	23.0	-54.8	62.0	51.8
24	23.0	-45.0	62.0	42.0	22.0	-50.5	63.0	47.5

* Adjusted to zero transmission level point

Table V AN/TRC-56 Multiplex
Echo Attenuation

REFERENCE:

Philco Record Book: 11729
Sheet: No. 46
Date: August 18, 1962

TEST CONDITIONS:

All channels set for 0 dbm in and -6 dbm out in 2-wire operation.

TEST RESULTS:

Channel Number	Unit A to Unit B (db)	Unit B to Unit A (db)
1	38.5	42.0
2	40.2	39.5
3	40.5	42.2
4	42.3	38.2
5	39.0	38.5
6	40.3	39.2
7	41.7	38.1
8	37.0	39.3
9	42.2	40.5
10	39.3	34.4
11	39.7	36.9
12	39.0	38.6

Table VI AN/TRC-56 Multiplex Dialing Distortion

REFERENCE:

Philco Record Book: 11729
Sheet: Nos. 42 and 43
Date: August 16, 1962

TEST RESULTS:

Channel Number	Unit A to Unit B t^1 (msec)	Unit A to Unit B Dist.* (%)	Unit B to Unit A t^1 (msec)	Unit B to Unit A Dist.* (%)
1	1.4	1.96	0.1	0.14
2	1.4	1.96	0.3	0.42
3	0.2	0.28	0.4	0.56
4	0.5	0.70	0.6	0.84
5	0.2	0.28	1.6	2.24
6	2.2	3.08	1.3	1.82
7	0.2	0.28	0.2	0.28
8	1.6	2.24	1.8	2.52
9	1.0	1.4	1.8	2.52
10	1.6	2.24	1.5	2.1
11	1.0	1.4	0.3	0.42
12	1.2	1.68	0.3	0.42

t^1 is the time variation of dialing pulses

* referenced to fourteen pps dialing rate

Table VII AN/TRC-56 Multiplex Channel Distortion

REFERENCE:

Philco Record Book: 11729
Sheet: No. 50
Date: August 21, 1962

TEST CONDITIONS:

Radio set and multiplex connected as a one-hop system.
0 dbm input, -3 dbm output; 2-wire operation;
channels 1 through 12
-16 dbm input, -3 dbm output on channels 22 and 24.

TEST RESULTS:

Channel Number	Unit A to Unit B			Unit B to Unit A		
	1 kc (db)	2 kc (db)	3 kc (db)	1 kc (db)	2 kc (db)	3 kc (db)
1	0	-50.0	-58.0	0	-48.0	60.0*
2	0	-45.0	-59.0	0	-44.0	-60.0*
3	0	-48.5	-55.5	0	-53.0	-60.0*
4	0	-51.5	-58.0	0	-53.5	-60.0
5	0	-52.0	-59.0	0	-55.5	-54.0*
6	0	-48.0	-59.0	0	-51.0	-60.0*
7	0	-50.0	-55.0	0	-47.0	-58.0
8	0	-47.5	-59.0	0	-46.0	-58.0
9	0	-49.0	-57.5	0	-53.0	-58.0
10	0	-45.0	-57.0	0	-53.0	-58.0
11	0	-43.5	-55.0	0	-56.0	-54.0
12	0	-42.5	-51.0	0	-54.0	-58.0*
22	0	-50.0	-53.0	0	-54.0	-59.0
24	0	-44.5	-56.0	0	-53.0	-59.0

* Limit of measuring technique

Table VIII AN/TRC-56 Multiplex Measured Crosstalk, A to B#

TEST CONDITIONS:

Radio set and multiplex connected as one-hop system.
0 dbm in, channels 1 through 12, 2-wire, -3 dbm out.
-16 dbm in, channels 22 and 24, -3 dbm out.

REFERENCE:

Philco Record Book: 11729
Sheet: No. 52
Date: August 21, 1962

TEST RESULTS: S/Crosstalk in db

		LOADED CHANNEL														Limit of
		1	2	3	4	5	6	7	8	9	10	11	12	22	24	Measuring Capability
M	1	0	73	76	66	71	76	76	70	69	69	72	60	60	73	66
E	2	55*	0	56*	55*	53*	53*	58*	66*	68	67	63	68	67	72	72
A	3	70	70	0	69	68*	72	71	72	72	72	72	72	71	72	71
S	4	69	70	66	0	70	69	68	69	69	72	70	71	67	71	71
U	5	55*	57	56*	56*	0	58*	61*	56*	57*	59*	56*	58*	63*	64*	65
R	6	68	65	68	65	66	0	69	69	67	67	66	67	67	69	71
E	7	70	70	70	70	72	70	0	69	72	70	72	73	67	72	72
D	8	66	67	67	69	67	58*	56*	0	49*	49*	55*	49*	51[1]	52*	52
	9	70	71	70	69	69	70	70	71	0	69	70	72	72	69	73
C	10	55*	51*	52*	56*	59*	69	69	69	67	0	67	66	55[1]	53[1]	69
H	11	69	58[1]	68	68	70	67	64	65	59[1]	66	0	61	64	61	61
A	12	51[1]	66	66	68	66	68	68	64	67	56[1]	56	0	53[1]	60	68
N	22	52[1]	59*	56*	59*	60*	56*	58*	58*	59*	59*	58*	57[1]	0	60	61
N	24	54*	55*	59*	59*	62	67*	63	62*	67*	67	67	56[1]	62	0	67
E																
L																

* Denotes spurious pickup (AM broadcast stations)
1 Denotes reading which is out of specification
A to B is an arbitrary unit designation

Table IX AN/TRC-56 Multiplex
Measured Crosstalk, B To A

TEST CONDITIONS:

Radio set and multiplex connected in one-hop system.
0 dbm in, channels 1 through 12, 2-wire, -3 dbm out.
-16 dbm in, channels 22 and 24, -3 dbm out

REFERENCE:

Philco Record Book: 11729
Sheet: No. 53
Date: August 22, 1962

TEST RESULTS: S/Crosstalk in db

MEASURED CHANNEL	LOADED CHANNEL 1	2	3	4	5	6	7	8	9	10	11	12	22	24	Limit of Measuring Capability
1	0	69	68	72	71	72	60	64	61*	70	70	64	74	72	76
2	69	0	63*	67	71*	67	63	63	67	66*	56[1]	65*	64	70	68
3	63	71*	0	66	74	71	68	66	66	71	74*	74*	64	60*	66
4	76	68	67	0	68	68	68	68	67	69*	64	69	69	74*	68
5	64	76	70	63*	0	68	74*	74*	71	72	63	72	70	68	76
6	62	64	67	66	64	0	69	63*	61	62	66	63	69*	67	71
7	74	72	71	74	74*	74	0	70	64	63*	61	72	70	71	74
8	68	65	68*	67	71*	69	67	0	71*	62	68	62*	59[1]	69*	74
9	73	67	64*	66	74	76*	65	74*	0	68	67	74	72*	70*	64
10	69	61	65	71*	65	63*	67*	65	66	0	66	67	68	61	69
11	69	69	68	71	69*	62*	64*	68	74*	66*	0	72	74	66	74
12	65	62*	63	68*	65	60	60	60	59*	58[1]	64	0	69*	63	69
22	59[1]	64	65	67	68	65*	67	60*	61*	64	63*	61*	0	61*	66
24	68*	68*	60*	65	62*	69*	70*	67	61*	69		64	67	0	70

* Denotes spurious pickup (AM broadcast stations)

1 Denotes reading which is out of specification

2. Functional Description

a. General

The Transmit-Receive Assembly has two basic functions: (1) the amplitude modulation of channel information on a given carrier, and (2) the selection and demodulation of the information being sent over the channel from the received baseband signal. Requirements for low crosstalk, flat frequency response, and low harmonic distortion determined the circuit complexity.

b. Transmit Function

The transmitter isolation transformer provides a balanced-line input. Resistor R150 is used to adjust the audio input level so that different nominal inputs can be adjusted for 30-percent modulation at the reference modulator, after amplification in the adjustable-gain amplifier. (See Figures 39 and 40.)

The reference modulator consists of transistors Q4 and Q5, and is driven by an emitter follower which provides high input impedance and isolation for the keying circuit. With no signaling or dialing voltage being applied, diodes CR1 and CR2 are reverse-biased. The positive voltage of the voltage divider, R22 and R23, and the bias voltage of Q4 bias the diodes.

During signaling, the positive keying voltage biases the diodes in a forward direction, a-c grounding Q4 base and also saturating Q4.

The reference modulator output passes through FL1, a bandpass filter centered at 150 kc. The output signal of FL1 is amplified in the transmitter reference frequency amplifier which drives the transmitter channel modulator. The channel carrier amplifier provides both channel modulators with a large carrier signal to switch the modulator diodes. The output of the transmit channel modulator is filtered by FL2. FL2 is a bandpass filter whose center frequency is that of the channel frequency. The common termination pad matches filter FL2 and provides isolation between channels.

c. Receive Function

The combined received baseband signal is filtered by FL3, which selects the desired channel frequency. This signal is then amplified by the three stages in the receiver channel frequency amplifier which has two negative feedback loops to extend frequency response and ensure stability.

The receive channel modulator is the same as the transmitter channel modulator. The product outputs of the channel modulator are filtered by FL4 which is a 150-kc bandpass filter. The receiver reference frequency amplifier enlarges the signal and drives a signaling driver and the envelope detector. The signaling driver is used in detecting carrier "on and off" and feeds the Signaling and Termination unit. The envelope detector consists of two diodes and a two-stage amplifier in a negative feedback configuration which provides a very low-distortion detector. An emitter follower matches the detector output to the low-pass filter. An adjustable-gain, three-stage, negative-feedback amplifier is used as the output audio amplifier. The audio output is connected to balanced lines through an isolation transformer.

3. Design Approach

a. General

Channel performance of the multiplex is determined largely by the Transmit-Receive Assembly. The specifications which influenced the over-all design considerations most were crosstalk, frequency response, level inputs and outputs, harmonic distortion, and temperature. Signal to noise, envelope delay distortion, and other specifications had less effect on the definition of the functional blocks.

b. Initial Design Approach

The initial design approach used a double-modulation process in the transmitter, with the audio amplitude modulating a 2500-kc carrier. This process was to be repeated on each channel. Harmonics of 2500 kc would be attenuated easily by a bandpass filter with a skirt selectivity such that audio distortion products which fell outside the 50-kc band allotted to each channel would be well attenuated. A second modulation process then would mix the 2500 kc with carriers varied in 50-kc steps to produce the desired amplitude modulated channel carrier. This approach rendered no undesirable products which could not be easily filtered out, assuming good suppression of carrier by the channel modulator.

Several problems were encountered which made the scheme unpractical. Balance of a channel modulator at 2.5 Mc was found to be difficult. Radiation at these frequencies demanded very efficient shielding to prevent crosstalk. This would contribute much to the over-all weight. Manufacture of the desired filters was not feasible.

c. Final Design Approach

The advantages of the double-modulation scheme were still to be employed, but the reference frequency had to be made lower. A reference frequency of 150 kc was chosen, since a bandpass filter with sharp skirts could be built at that frequency and the harmonics of 150 kc could be easily attenuated. The effort on the transmitter of the multiplex entailed design of an adjustable-gain amplifier, a reference modulator which could be keyed on and off, and a channel modulator which balanced out the carrier and the input. A double-modulation scheme similar to that in the transmitter of the multiplex would also be employed in the multiplex receiver. The audio was to be obtained with synchronous detector which theoretically would be capable of zero-percent distortion. A double-modulation scheme was necessary to bring down the carrier frequency, since control of the phase of the sampling carrier would be difficult over the entire channel spectrum. Synchronous detection would take place at 150 kc, and the sampling carrier would be obtained from the signal being detected.

Figure 41 is a simplified block diagram of the intended Transmitter and Receiver.

The 150-kc carrier from the synchronous detector would be used for signaling purposes.

Frequency response specifications were to be met by apportioning to the different blocks suitable response specifications. These were chosen such that, in the worst case, with the responses adding in the same direction the overall response would still be within the total system specification. Little contribution of ripples and droops in the response was expected from the amplifiers or modulators. Primarily, the frequency response was dictated by that of the filters within a channel. Envelope delay distortion was to be avoided by making all frequency responses as flat as possible and extending them beyond the pass band, where feasible, to avoid drastic rate of changes in the phase response on the edges of the pass band.

4. Results

The systems test results given in Section A3, in many cases, represent the total performance of the Transmit-Receive Assembly. Results considered in this section will not cover those already illustrated, but will cover the individual performance of the more important functional blocks on the Transmit-Receive Assembly.

The reference modulator used in the transmitter is a carrier-driven switch which samples the audio signal at a 150-kc rate. Such a modulator inherently produces low audio distortion sidebands about the carrier frequency.

An equivalent circuit of the modulator is shown in Figure 42. The switch representing the transistor operates at the carrier rate, alternately connecting the input audio to ground through 100 ohms and 680 ohms. Although the modulator has very low second- and third-harmonic sidebands around the 150-kc carrier, the output wave is rich in carrier harmonics. However, the 150-kc bandpass filter following the modulator attenuates these to 60 db below the first-order sidebands about the 150-kc carrier.

Another equivalent circuit which illustrates more clearly the modulation process is shown in Figure 43. The combined effect of the audio signal and B+ on the switching transistor is shown as a battery which varies in voltage at an audio rate. For 100-percent modulation the battery varies in voltage from 2 B+ to 0, being 2 B+ at the point where the peak audio is positive, and 0 at the point where the peak audio is negative. These two points are indicated as (1) and (2), respectively, on the output waveform. For the case of no audio input the battery is constant at B+ and the output consists of a train of equal-amplitude pulses occurring at the carrier rate.

Observation of the output waveform shown in Figure 43 shows that the modulator multiplies the audio signal alternately by "1" and "0"; "1" when the switch is open (transistor cut off), and "0" when the switch is closed (transistor saturated). To mathematically describe the modulator it becomes necessary to find a mathematical expression describing the switching function and then multiply this function by the input audio.

The mathematical expression for the switching function is obtained by deriving the Fourier series expansion for a continuous train of rectangular pulses having a peak pulse amplitude of unity and a repetition rate equal to the carrier frequency. The switching function is found to be:

$$U(t) = k \sum_{n=-\infty}^{\infty} \frac{\sin(nk\pi)}{nk\pi} \cos(n\omega_c t) \qquad (1)$$

where: $k = t_o/T$ (duty cycle)

t_o = time duration of the pulse in seconds

T = time period between pulse centers in seconds

$\omega_c = 2 f_c$ (angular carrier frequency in radians per second)

Assuming a sinusoidal audio signal, the audio input to the modulator is given by:

$$E_s = A_m \cos(\omega_m t + \theta_m) + C \tag{2}$$

where: A_m = peak amplitude of audio signal

ω_m = angular frequency of audio signal in radians per second

θ_m = arbitrary phase angle of audio signal

C = effective B+ (d-c) input to modulator

Multiplying the switching function (Equation 1) by the audio input signal (Equation 2) gives:

$$E_o = kA_m \cos(\omega_m t - \theta_m) + kC + 2\,Ck \sum_{n=1}^{\infty} \frac{\sin(nk\pi)\cos}{nk\,\pi} (n\omega_c t) + kA_m \sum_{n=1}^{\infty} \frac{\sin(nk\,\pi)\cos}{nk\,\pi} \left[n\omega_c t \pm (\omega_m t + \theta_m)\right] \tag{3}$$

In Equation 3, the first term represents an audio output; the second term, a d-c output; the third term, carrier signals at the fundamental and harmonic frequencies of the switching carrier; and the fourth term, audio sidebands on either side of the fundamental switching carrier and its harmonics.

It is noteworthy that for perfect square wave switching, i.e., k equal to 1/2, all terms involving even values of n in the infinite series of Equation 3 vanish. This means that the modulator output contains carrier signals with audio sidebands only at the fundamental switching carrier and its odd harmonics. This output spectra is shown in Figure 44, where the carrier frequency is 150 kc and the audio signal frequency is 20 kc.

Since the switching function is not an ideal square wave, a typical output spectrum does contain some low-level even harmonics.

Frequency responses of typical 150-kc bandpass filters are shown in Figures 45 and 46, illustrating that harmonics of 150 kc are attenuated by more than 70 db.

Two of the primary concerns in the development of the channel modulator were distortion and crosstalk. The function of the channel modulator is to place the 150-kc modulated wave from the 150-kc bandpass filter into its proper frequency slot, as shown in Figure 47.

Figure 48 illustrates the crosstalk problem which could occur if the channel modulator were improperly designed and simultaneously (a) failed to suppress the modulating signal; (b) failed to suppress the channel carrier and its harmonics; and (c) distorted the modulating signal and modulated these distortion products onto the carrier and its harmonics. In Figure 48, f_{c1} is the 150-kc carrier, f_a is the audio signal modulating f_{c1}, and f_{c2} is the channel carrier driving the channel modulator (assumed to be 200 kc for the example). Harmonics of the modulating signal were assumed to follow a 1/n curve, where n is the harmonic number. In Figure 48, signals which overlap the areas 50 kc ±20 kc and 350 kc ±20 kc represent crosstalk signals which cannot be eliminated by filters.

Many balanced modulators were studied, but the ring modulator was found to be most satisfactory. Figure 49 shows the two switching states of the ring modulator in a simplified drawing. Assuming the modulator diodes to be perfect carrier-driven switches and the modulator to be perfectly balanced, the switching function performed by the modulator is described by the following expression:

$$\dot{U}(t) = \sum_{n=-\infty}^{\infty} \frac{\sin(n\pi/2)}{n\pi/2} \cos(n\omega_{c2}t) - 1 \tag{4}$$

where ω_{c2} = angular frequency of the channel carrier in radians/second.

The output of the 150-kc bandpass filter is the modulating signal input to the channel modulator and is given by;

$$E_m = A_c \cos(\omega_{c1}t) + A_m \cos\left[\omega_{c1}t \pm \omega_m t\right]$$

where: A_c = peak amplitude of 150-kc carrier

A_m = peak amplitude of audio sidebands about 150-kc carrier

ω_{c1} = 150-kc carrier frequency expressed in radians per second

ω_m = angular frequency of audio signal in radians per seconds

Multiplying Equations 4 and 5 together results in the following channel modulator output:

$$E_o = A_c \sum_{n=1}^{\infty} \frac{\sin(n\ \pi/2)}{n\ \pi/2} \cos\ (n\omega_{c2}t + \omega_{c1}t) +$$

$$A_m \sum_{n=1}^{\infty} \frac{\sin(n\ \pi/2)}{n\ \pi/2} \cos\left[n\ \omega_{c2}t \pm (\omega_{c1}t \pm \omega_m t)\right]$$

Figure 50 illustrates this output for $n < 3$, $A_c = 1$, $A_m = 0.5$, and $\omega_{c2} = 2\ \pi(200\ \text{kc})$.

Bandpass characteristics of a Voice and Data channel filter are shown in Figures 51 through 54. Channel separation resulting from the channel filters is seen in Figure 55. The effect of the double-modulation process can be seen by the channel separation achieved as a result of the sharper bandpass of the 150-kc filters. Figure 56 shows a plot of three typical 150-kc bandpass characteristics removed 50 kc from one another.

In the receiver, the channel filter, channel modulator, and the 150-kc bandpass filter are identical to those in the transmitter. Sufficient level flexibility, gain, and attenuation tolerances through a channel eliminate the requirement for adjustment of levels on any channel at baseband frequencies. Demodulation of the AM modulated 150-kc carrier was found to be impractical with a synchronous detector because of reproducibility problems. A novel circuit

was used to envelope detect the signal. The circuit was piecewise linear as opposed to the continuous nonlinearity of most envelope detectors at small signal levels.

Figure 57 shows a standard diode curve and that of the envelope detector used.

Performance of the envelope detector was extremely satisfactory. For the nominal level input at 30-percent modulation, the worst-case second harmonic distortion throughout temperature range was 0.25 percent. Potentially, the envelope detector was capable of even less distortion; however, further development was not necessary.

The envelope detector was followed by a matching stage and a low-pass filter. Figures 58 and 59 show the responses of the two low-pass filters used. The 4-kc flat filter is used on voice channels, and the 20-kc flat filter is used on the data channels. The channel response is primarily limited by the low-pass filters. The 150-kc bandpass filters are made essentially flat for 20-kc data usage. . The channel filters are flat within 1 db ± 15 kc about the center frequency for voice channels, and flat within 0.3 db ± 20 kc about the center frequency for data channels.

The dropout audio amplifier provides an adjustable high-gain amplifier which can provide large levels of audio.

5. Practical Considerations

The overall performance of the Transmit-Receive Assembly is good. However, to ensure good performance, the circuitry is perhaps too sophisticated. Elimination of the double modulation scheme on the transmit side would be possible if a low-distortion reference modulator could be built at all channel frequencies. The channel separation that was achieved was far more than what was found to be required. The reference modulator used, when tested at higher frequencies, inserted out-of-phase carrier which results in distortion when the signal is envelope detected. These considerations should be of prime interest from the weight aspect and from the aspect of simplifying the unit. In the receiver, a further reduction of circuits could be accomplished by amplifying the total baseband signal received instead of amplification on a per-channel basis. This, of course, would require design of a receive baseband amplifier which had very good dynamic range. Further considerations could be given to the method of combining the baseband output signals of the transmitters. The technique used attenuated the signal to such a low level that interference signals from the AM

broadcast band were picked up. An improved technique would also improve the signal-to-noise ratio of the channels, since it was at this low signal level point that the signal-to-noise ratio was determined by the first transistor on the baseband amplifier.

C. SIGNALING AND TERMINATION ASSEMBLIES

1. Mechanical Description

The Signaling and Termination Assemblies are located on 10 by 4.5 inch printed boards. The switches which are used to select the different options are placed on mechanical brackets and are located so as to economize on board space. It was because of this reason that the switches were not located for front access. A Signaling and Termination patch board is provided for trouble shooting. The Signaling and Termination patch board is mounted on the Transmit and Receive patch board when not in use. Connectors on the Signaling and Termination Assembly are the same as those on the Transmit-Receive Assembly.

2. Functional Description

The block diagrams of Figures 63 and 62 show the functional blocks in the Signaling and Termination Assembly for two modes of operation; AC Ringdown and E & M Dialing. Figure 61 shows the schematic of the Signaling and Termination Assembly. The connection or disconnection of certain blocks is made by operating the switches on the printed board. Switches 1 and 2 predispose the Pulse Orientator for eight different types of E & M dialing pulses. With the dialing input connected to the 24-volt or 48-volt input and the switches in their proper positions, the Pulse Orientator delivers a 13-volt peak-to-peak square wave to the keying circuit which turns the reference modulator on and off at the dialing rate. In the Pulse Orientator circuit, a zener diode of 6.2 volts prevents false signaling from occurring. Table X shows the different types of dialing and switch positions for each. As an example, for the pulse shown in Figure 60, the carrier should be on and off accordingly.

Table X indicates that switches 1 and 2 are in position 1. The +24 volts are connected to the emitter through a resistor which turns off transistor Q1. (See Figures 61 and 62.) Transistor Q2 is then biased on and turns the carrier off at the keyer. When the phone is off the hook, Q1 is on and Q2 is biased off, turning the carrier on. The carrier then is turned off and on at the dialing rate. After the dialing sequence has ended, the carrier remains on to carry information. The carrier is turned off when the phone is replaced on the hook.

Table X Switch Positions for
E & M Dialing

	Plus to Ground	Ground to Plus	Minus to Ground	Ground to Minus	
24V	2	1	1	2	SW 1
48V	2	1	1	2	
24V	2	1	2	1	SW 2
48V	2	1	2	1	

On the receive side of the Transmit-Receive card, the 150-kc carrier triggers the signaling driver circuit on and off. Ideally, the signaling driver recognizes only the presence or absence of carrier and is not affected by normal modulation. At the input of the receive side of the Signaling and Termination Assembly, two diodes in parallel and a parallel tuned circuit prevent false signaling. The first driver amplifies the 150 kc and drives a low-frequency envelope detector which removes the signaling envelope. For E & M dialing operation switches Sw 4 and Sw 5 are placed in the Dial position. This connects the base of Q3 to the emitter of Q5, and the E & M relay to the Q3 collector. The signaling voltage at the emitter of Q5 turns Q3 on and off, which energizes and de-energizes the E & M relay at the dialing rate.

In AC Ringdown operation the signaling sampler pulses the Pulse Orientator on and off at the signaling rate. (See Figure 63.) The signaling sampler is inoperative for all voltages less than 12 volts peak. For AC Ringdown operation the Pulse Orientator switches 1 and 2 are placed in positions 2 and 1 respectively, and switches 4 and 5 are placed in Ring position. At the receive side, the carrier on and off is detected at the envelope detector; however, after passing through Q5 the envelope is rectified and used to turn on the AC Ringdown relay which energizes the 2-wire line with 20 cps from the ringing supply. Switches Sw 4 and Sw 5 in the Ring position connect the base of Q3 to the emitter of Q4, and the AC ringdown relay to B+20 volts and the collector of Q3. The double-detection method is necessary to prevent a carrier loss from turning on the AC Ringdown relay.

Four-wire to 2-wire conversion is accomplished with a resistive hybrid. The hybrid was designed for a 16-db loss in the transmit path and 2 db of attenuation in the receive path. A variable resistor, R16, is used to balance the hybrid to the 2-wire line impedance. Coil L2 balances out capacitor C2 in the 2-wire line. Switch Sw 3 provides the option of disconnecting the transmit path from the hybrid so that a 4-wire configuration is made available. The 2-wire line input now becomes the receive path of the 4-wire line. The signaling squelch consists of two circuits which protect the transformers and their immediate circuitry from the large AC Ringing voltage.

3. Design Goals

The basic design goal of the Signaling and Termination Assembly was to provide a unit which would work with the different standard signaling and dialing methods employed. The different options were to be provided with no degradation of the frequency response, distortion, etc. Balance of the 2-wire to 4-wire hybrid had to be such that echo attenuation through the system was 25 db or greater. The method of signaling to be used was the turning on and off of the channel carrier at the signaling or dialing rate.

The problem then was to design circuitry to program the dialing or signaling voltages so that the carrier went off and on in a proper sequence, and that the reception of the on-off carrier either applied a 20 cps ringing voltage to the 2-wire line or turned a relay switch on and off at the dialing rate. Circuitry was also contemplated which would prevent oscillation of the system, when the hybrids were unbalanced, by squelching the receive side from transmitting while signaling was taking place; and also circuitry which would protect the 4-wire inputs of the Transmit-Receive Assembly from the large signaling voltages.

4. Results

All the design objectives were met except for receive-end squelching as described in the previous section, and satisfactory operation of the protective circuitry. With the hybrids balanced to 600 ohms and the receive and transmit levels set to -6 dbm and 0 dbm respectively, the worst case of echo attenuation was 34.4 db. See Table V for complete data.

An attempt was made to measure harmonic distortion in and out of the 2-wire line; however, the readings were found to be negligibly small. The frequency response, using the 2-wire line, was almost flat throughout the pass band and contributed only a slight droop at the low-frequency end. Signaling and dialing were successfully accomplished. The E & M dialing distortion values are given in Table VI. Although it is not specified, the dialing distortion was within common telephone practice specifications. The Signaling Squelch circuit was put in to protect the 4-wire input and output of the Transmit-Receive Assembly. Relay K3 was to open the 2-wire line when large signaling voltages were present so that the on-off carrier would not be modulated during its on cycle by the ringing voltage which was divided down by C2 and zener diodes CR13 and CR14. However, during systems test, it was found that relay K3 was not opening during signaling. The investigation was not pursued to any length because there was no adverse contributions to system performance by the carrier being modulated during its on cycle. The Signaling and Termination Assembly satisfactorily met its design objectives aside from the considerations discussed.

D. CARRIER GENERATOR ASSEMBLY

1. Mechanical Description

The Carrier Generator Assembly is contained in a rectangular box in which are located 14 printed cards. Connection is made to the various

Transmit-Receive Assemblies through threaded coaxial connectors. A terminal board is used on the exterior of the package to connect to +20V and ground. There are two banks of cards inside the package, held in place by the removable top and bottom plates. The top bank of cards are the harmonic generator and the six 3-stage filters. The bottom bank of cards are the 2-stage filters. Metal tracks on the inside of the package, and the top and bottom plates hold the cards securely in place.

2. Functional Description

The Carrier Generator Assembly provides two groups of carriers for modulation purposes. Mixed with the 150 kc, the upper and lower sideband products about the low-frequency group produce all the channel carriers from 50 kc to 600 kc, in 50-kc intervals. The high-frequency group of carriers mixed with 150 kc produce all channel carriers from 650 kc to 1.2 Mc. Table I shows the frequency plan for a 24-channel multiplex. All the carriers are derived from the harmonic generator card which consists of a 100-kc, crystal-controlled oscillator and a 2-for-1 divider network. (See Figures 64 and 65.) The 50-kc harmonic generator circuit is driven by the 100-kc signal from the oscillator and produces two identical waveforms which are rich in harmonics. One output drives the 2-stage, low-frequency filter cards, the other drives the 3-stage, high-frequency filter cards. The 2-stage filter cards consist basically of two common emitter stages with series tuned circuits connected from emitter to ground. The series tuned L-C sections are tuned to the desired frequency by an adjustable capacitor. At resonance the emitter circuit has very low impedance and the amplifier stage has maximum gain. In this fashion, filtering and amplification are integrated into one operation. The 2-stage and 3-stage filter cards are all capable of producing 1-volt peak-to-peak carriers. The 3-stage active filters are the same as the 2-stage except for an additional stage and an AGC circuit. The additional stage is necessary because the higher harmonics are lower in level and the filtering Q has to increase as the frequency becomes higher. The AGC circuit was included to provide a stable amplitude output with respect to temperature.

3. Design Goals

The design goals for the Carrier Generator were dictated by the frequency plan and the type of modulators to be used. To avoid routing of large-level signals, the carrier generator outputs were set to be 1 volt peak-to-peak. Further amplification is made on the Transmit-Receive Assembly. The basic requirements of the Carrier Generator outputs are amplitude stability within 3 db and frequency stability within plus or minus 2 cps. To meet the crosstalk specifications, adjacent carriers are attenuated 50 db.

4. Results

A suitable waveform was obtained from a divider circuit which was rich in harmonics. The harmonics of the fundamental were found to decrease in amplitude at a $\frac{1}{n}$ rate where n is the number of the harmonic. Figure 66 shows the spectral distribution of the waveform. The 100-kc crystal oscillator provides frequency stability which does not meet specifications. Undoubtedly, the stability could be improved; however, the real problem of carriers being different in a transmitter and receiver is the production of unwanted tones in the channel. The frequency stability of the units is such that in the worst case the tones would be audible; but with FIA weighting, the quality of the channel would still be considered in a qualitative fashion as being good. Worst case presupposes one equipment being in the highest temperature environment, the other equipment being subject to the lowest temperature environment, and the 100-kc crystals at their worst possible tolerances.

The outputs of the filters are adjustable to 1-volt peak-to-peak and are amplitude stable within 3 db over the entire temperature range. The AGC circuits were found necessary to keep stable the amplitudes of the higher frequency cards. The amplitude changes without the AGC were due to the change of Q in the L-C tuned stages with temperature.

The attenuation of adjacent carriers in the worst case measured for two Carrier Generators was 54 db. Thus, aside from frequency stability, the Carrier Generator met its design goals. The basic idea of producing a rich harmonic waveform and the subsequent filtering allows all the carriers to be integrally related in frequency, and is a relatively simple approach to the problem of carrier generation.

E. POWER SUPPLY ASSEMBLY

1. Mechanical Description

The Power Supply Assembly is an integral unit on which are mounted the front panel and three circuit boards. The three circuit boards are the baseband amplifier, the power supply components boards, and the metering and test board. The front panel of the Power Supply Assembly contains components which provide: adjustment of the 20-volt d-c level, the 1-kc oscillator output, and the baseband amplifier level output; switching of the equipment on or off and selection of 400 cps or 50/60 cps operation; d-c and a-c power indications; metering circuit and 1-kc oscillator connections; and switchable metering functions.

The baseband amplifier is wholly contained on the printed board. Room for an identical amplifier was left on half of the board. Small components of the power supply were placed on a circuit board. These components are connected by point-to-point wiring. Metering and Test Facility components not mounted on the front panel are located on the Metering and Test Facility printed boards. The heavier components of the power supply are located on the bottom shelf of the assembly to lower the center of gravity. Transistors used in the voltage regulator, electronic filter, and 400 cps inverter, which handle relatively large currents, are placed inside a blower duct mounting assembly on forced-air, convection-type heat sinks. This duct is continued on the bottom back of the equipment and guides the air flow such that the printed card assemblies receive uniform cooling.

2. Metering and Test Facility

a. Functional Description

The Metering and Test Facility provides a 1-kc signal to enable adjustment of levels and the balancing of hybrids in a channel. The 1-kc oscillator consists of a free-running multivibrator whose frequency is controlled by a resistor-capacitor feedback combination. (See Figures 67 and 68.) The 1-kc signal is fed into an emitter follower which drives a low-pass filter, eliminating the higher-order harmonics. An adjustable-gain amplifier after the filter provides a variable amplitude output. This output can be connected to the inputs of a channel with a patch cord provided with the equipment.

The Function Switch, S3, of the Metering and Test Facility interconnects the circuitry so that different levels of signal and d-c voltages can be read on the meter scale. In positions 1 and 2, J1, the Balanced Input, is connected through an isolation transformer to either of two matching pads. The matching pad is connected to an adjustable-gain amplifier whose gain is preset to give proper meter accuracy. The two matching pads used in positions 1 and 2 have 600-ohm inputs, and one has 10 db more attenuation. The d-c rectifier which follows the adjustable-gain amplifier converts all a-c signals to d-c. The d-c microampere meter is protected by a diode so that large levels would not cause meter failure. The resistor configuration between the d-c rectifier and the meter is necessary to ensure meter accuracy over the temperature range. In positions 3 and 5,J2 is connected to the amplifier. These two connections provide a sufficiently high impedance so that the points being monitored are not loaded. Position 4 is a meter-off condition, and positions 6 and 7 connect the meter to the two d-c voltages used in the equipment, through padding resistors.

b. Design Goals

The design of the Metering and Test Facility is a result of the equipment specification requirement that all voltages requiring adjustment during operation shall be read with an integral meter.

The equipment is designed for unattended operation after the initial adjustments have been made. Periodic checks should be made only to guarantee that failure or deterioration of a component has not degraded the performance of the unit. The Metering and Test Facility was designed to aid in the initial adjusting of the equipment and the monitoring of the more important voltage points.

c. Results

The functions provided by the Metering and Test Facility include the following:

(1) Monitoring of the 20 VDC and 6.8 VDC levels to check the performance of the power supply.

(2) Setting of a nominal audio input level to the proper modulation index.

(3) Balancing of the hybrids.

(4) Setting the dropout amplifier on the Transmit-Receive Assembly to a given nominal output level.

(5) Measuring of the carrier levels.

Range of the metering could be extended to allow setting of a wide range of audio levels in and out if some redesign were to be done. However, the metering is sufficient to set the equipment to standard operating levels. The frequency of the 1-kc oscillator used in the initial adjustments is not rigidly controlled, since the pass band is very flat. One monitoring facility of interest which was not provided was the measuring of baseband in and out signal levels. There is no simple approach to this task because the baseband signal is a composite of many channel frequencies. Schemes which involve switching all channels out except one require cumbersome circuitry and interruption of communications. These facts make it undesirable.

3. Baseband Amplifier

a. Functional Description

The Baseband Amplifier consists of three common-emitter amplifier stages to amplify the combined baseband signal output of the channels. (See Figure 69.) All three stages have some feedback in each emitter. The two main feedback loops extend the frequency response, with the feedback loop from Q2 emitter to Q3 emitter lowering the output impedance of the amplifier to 75 ohms for proper matching.

b. Design Goals

The purpose of the Baseband Amplifier was to provide suitable amplification of the combined baseband output signal before being connected to the radio unit. The basic design goals were a 20-db gain amplifier with a frequency response extending past 1.2 Mc and low intermodulation distortion.

c. Results

The amplifier designed was basically the same as the baseband amplifier in the Radio Receiver section, with some modifications. Figures 70 and 71 show frequency response and noise-to-noise readings respectively of a typical unit. Gain of the amplifier is 0 to 20 db, adjustable.

4. Power Supply

a. Functional Description

The input to the multiplex Power Supply can be connected to a 400-, a 50-, or a 60-cps, 117-volt a-c source. (Refer to Figure 72 and the block diagram on Figure 73.) Switch S1 turns the supply on and off. DS1 is a front panel light which indicates that the power is on. CR1 is a transient suppressor to protect the circuitry. Switch S2 connects either of two transformer taps to the input, depending on the source frequency. Transformer T1 has two output windings.

Output Winding 6 - 7 supplies a-c voltage to a rectifier bridge. The d-c voltage is filtered by capacitor C6 and drives the 400-cps inverter circuit. The 400-cps voltage at the output of transformer T2 powers blower *B1.

Output Winding 4 - 5 supplies a-c voltage to the main d-c supply. In the d-c supply the a-c voltage is rectified in the full-wave diode bridge rectifier and filtered by capacitor C1. Transistors Q1, Q2, and Q3 are the active components of the electronic filter which eliminate most of the remaining ripple. The B+20V level is adjusted by potentiometer R18. A change of voltage at the base of Q12 causes a current change through Q11, which turns Q10 more on or off. The collector of Q10 changes voltage, altering the bias on transistors Q4, Q5, and Q6 through Q7 and Q8, to regulate the d-c line voltage. The short-circuit protection biases the voltage regulating transistor bank to an open condition, to prevent failures due to accidental shorting of the d-c output line. The 6.8-volt d-c output is supplied by a voltage-dropping resistor and a 6.8-volt d-c zener diode.

b. Design Goals

The Power Supply was to provide the d-c voltages to power all the circuits in the multiplex unit. Temperature specifications and the wide range of the power-source frequency were the important considerations in the design of the unit. Regulation of the d-c voltages was dictated by the type of circuits in the multiplex. Ripple and hum had to be low because of the large gain in the receive leg of the channels. Also, the power supply had to drive a 400-cps blower to cool the transistors, to extend their life.

c. Results

In the final testing of the completed units, the ripple was approximately 3 millivolts. Changes of a-c line voltage from 108 volts to 132 volts, at line frequencies of 400 cps and 60 cps, resulted in a maximum change in the 20-volt d-c output of 0.04 volt d-c in the worst case. With changes in temperature, the prototype unit 20-volt d-c output did not vary significantly, such that the circuitry in the multiplex would be affected. Changes at temperatures which were at least 10 degrees more extreme than those which would be encountered according to specifications were less than 0.45 volt d-c for the 20-volt d-c output.

SECTION III

RADIO

A. GENERAL

The AN/TRC-56 Radio is a small, portable, line-of-sight, microwave equipment designed for field use under extreme environmental conditions. It is convertible for operation in either of two microwave bands; one from 7.125 gc to 8.4 gc, the other from 14.0 gc to 15.4 gc. Its mechanical configuration and operating characteristics make this unit suitable for use where rapid establishment of communications is essential. A photograph of the radio set is shown in Figure 74.

1. Mechanical Description

The AN/TRC-56 Radio is constructed to withstand rugged handling and environmental extremes. The unit is transported in two parts: the tripod, and the radio package. The radio is equipped with a set of shock frames which are used during transit; for operation, the shock frames are removed and the radio is placed on the tripod. The tripod permits horizontal and vertical scanning, and contains a leveling mechanism and azimuth and elevation calibrations.

The side of the radio unit containing all operational controls is protected by a metal cover. This cover is used during transit or unattended operations; however, during equipment adjustment or attended operation the cover is removed. The control panel consists of two parts: the upper section, hinged at the top, may be opened to permit access to radio assemblies; the lower section is a part of the removable power supply. The controls pertaining to the power supply are located on the lower part of the control panel; all other controls are located on the upper part.

The radio consists of a parabolic antenna with replaceable feed, waveguide assemblies, transmit and receive klystrons, the various circuit assemblies, and a power-supply assembly.

The antenna feed, the klystrons, and the waveguide assemblies are interchangeable with corresponding units for either the low- or high-frequency band of operation. The other assemblies are common to both frequency bands.

All circuits, with the exception of the two klystrons, are entirely solid state. High-quality components, well derated for operation, are used throughout. The unit is cooled by two blowers, one mounted in the power-supply assembly, the other near the two klystrons. Protective thermostats, set to shut off the power in case of excessive temperature rise, are incorporated. One is located at the power supply, the other at the klystron; these points are the most critical in the unit.

All circuits, except the i-f preamplifier and the i-f amplifier, are constructed on glass-epoxy boards with point-to-point wiring. The i-f preamplifier is constructed on a metal chassis, while the i-f amplifier is constructed with replaceable printed-board stages. All units are sprayed for moisture and fungus resistance.

2. Functional Description

The basic block diagram of the radio unit is shown in Figure 75. The baseband signal is applied to a switch through the input connector. The switch, labeled XMTR TEST-OPR, is set normally to the operate position, and connects the baseband signal to the pre-emphasis circuit. The pre-emphasis circuit may be strap-wired for transmission of television signal, multiplex signal, or completely bypassed. The pre-emphasized signal is amplified by the transmit baseband amplifier and applied to the reflector of the transmitter klystron. The frequency of the transmitter klystron is deviated by the signal, resulting in FM modulation. The r-f signal from the klystron, passed through the frequency and power monitor and the isolator, is transmitted by the antenna.

Cross-polarization of the antenna feeds permits the antenna to be used for both transmission and reception. The preselector filter passes the received signal to the mixer-i-f-preamplifier chassis, where it is heterodyned to the 70 mc i-f frequency and amplified by 25 db. From the preamplifier the signal is passed through the i-f amplifier, where amplification, limiting, and detection take place. The i-f amplifier also contains squelch and received carrier level monitoring circuits. The baseband signal from the i-f amplifier is de-emphasized and amplified in the receive baseband amplifier. It is then connected to the output connector through a switch (RCVR TEST-OPR).

The d-c component from the output of the i-f discriminator is used for automatic frequency control of the local oscillator. This d-c voltage is applied to the AFC circuit, where it is amplified and applied in series with the constant power-supply voltage to the reflector of the local-

oscillator klystron. The frequency is adjusted to place the i-f signal in the center of the band, reducing the discriminator d-c component to a few millivolts. The AFC amplifier has sufficient gain and range to compensate for the normal frequency drift of both transmitter and receiver klystrons.

An order-wire system is incorporated in the radio to permit voice communication between the two radio sites. Voice-powered phones provide signal to amplitude-modulate the 8.5-Mc order-wire carrier, which is generated whenever the push-to-talk switch is depressed. When the push-to-signal switch is depressed, the order-wire circuit provides the 8.5-Mc carrier modulated by a 10-kc signaling tone. The order-wire carrier is inserted into the transmit baseband amplifier and transmitted.

On the receiver side of the radio, the baseband signal from the i-f amplifier is coupled to the order-wire receiver. This circuit, tuned to the order-wire carrier frequency, amplifies and demodulates the carrier, providing audio output to the phones.

For initial adjustment of system gain, the input and output switches are set to the Test position. A known reference signal is applied at the input, so that the output of the transmit amplifier may be monitored and its gain adjusted to the proper level. After this is done, the output of the receive baseband amplifier is monitored and its gain adjusted for proper system gain. Proper system gain depends upon the terminal equipment to which the radio is connected.

The radio unit also contains the power supply and a metering circuit. The power supply furnishes all voltages necessary for radio operation, and the metering circuits permit monitoring of various voltages to ensure proper operation.

In addition, an air-flow control circuit is provided. During high ambient temperature the air-flow control vanes are open and outside air is passed over the circuits for cooling. When the ambient temperature drops, the control vanes close and the cooling air stream is recirculated inside the unit. This system provides partial regulation of temperature inside the unit.

3. Design Goals and Test Results

The design goals for the radio unit were based on the AN/TRC-56 specifications, RADC 2562A, Amendment 7. Some of these are tabulated in Table XI, together with the results of the tests. The

tests were performed on two units, both in the laboratory and in the field. First, low-band r-f components were mounted into the units for testing; then these components were replaced for high-band operation. The frequencies used are as follows:

	UNIT 1		UNIT 2	
	Low Band (gc)	High Band (gc)	Low Band (gc)	High Band (gc)
Xmtr	8.350	14.100	7.210	15.300
L.O.	7.140	15.230	8.280	14.030

The tests were performed on individual units and also on the back-to-back connected system. No environmental system testing was performed; however, in the design stage, all of the subassemblies went through exhaustive temperature testing.

B. POWER SUPPLY

1. Mechanical Description

The radio power supply is built on a reinforced aluminum chassis. All of the larger components are bolted to this chassis directly; the smaller circuit components are mounted on two epoxy-glass component boards. The power transistors are mounted on a tubular heat-sink assembly, which is cooled with forced air. All controls necessary for operational adjustment are located on the front part of the chassis; the controls which require only one initial adjustment are located on a small panel on top of the unit.

The power-supply chassis slides into the radio case on two teflon strips. It is held in position by two heavy pins mounted on the radio case, and by screws removable from the outside. The front part of the power-supply chassis forms a part of the radio control panel.

2. Functional Description

a. General

The power supply is designed to operate from 117-volt, 50-, 60-, or 400-cps line, and to provide several d-c regulated outputs and a 117-volt, 400-cps unregulated output. The circuit diagram of the power supply is shown in Figure 86. Line voltage is applied to the input

Table XI Table of Specifications

Specification Item Number	Specification	Test Results
3.5.2.7	Transmitter	
3.5.2.7.1	The input impedance shall be 75 ohms resistive, unbalanced. The input level shall be from 0.5 to 2.0 volts peak.	Input Impedance: Unit 1, 77 ohms Unit 2, 81 ohms
3.5.2.7.2	The transmitter shall have a minimum power output of 1 watt throughout the lower frequency range, and 100 milliwatts throughout the higher frequency range.	Output Power: Low Band 990 mw at 7.21 gc 660 mw at 8.35 gc High Band 155 mw at 14.1 gc 280 mw at 15.3 gc
3.5.2.7.3	The minimum acceptable deviation ratio shall be 1.0.	1.1
3.5.2.8	Receiver	
3.5.2.8.1	The receiver shall have an output impedance of 75 ohms. The output level shall be adjustable from 0.5 to 2.0 volts peak.	Output Impedance: Unit 1, 93 ohms Unit 2, 87 ohms Output Adjustable from 0 to 2.5 volts peak.
3.5.2.8.2	The receiver noise figure shall be less than 11 db.	Noise Figure: Low Band 9.5 db Unit 1 10.5 db Unit 2 High Band 14.3 db Unit 1 12.3 db Unit 2

Table XI Table of Specifications (Continued)

Specification Item Number	Specification	Test Results
3.5.2.8.3	The receiver shall be equipped with AFC that shall have a pull-in range of at least 25 Mc.	AFC Pull-In Range: Low Band, ± 12.5 Mc High Band, ± 12.5 Mc
3.5.2.8.4	The i-f discriminator bandwidth shall be 20 Mc to the 3-db points; less than 30 Mc to the 20-db points; and less than 50 Mc to the 60-db points.	See Section III-E-4
3.5.2.8.5	The receiver shall have sufficient gain to produce the maximum output with full limiting for an input r-f signal within the threshold limitations of a receiver having the required bandwidth.	See Figures 76, 77
3.5.2.10	System Performance The following characteristics shall apply to the transmitter and receiver when connected back-to-back.	
3.5.2.10.1	The system shall have unity gain ± 5% in the input signal range of 1.0 to 2.0 volts.	Adjustable for unity gain.
3.5.2.10.2	The amplitude versus frequency response of the system shall be -0.5 db from 60 cps to 5.0 Mc; + 1 db, - 3 db from 20 cps to 7.5 Mc when referenced to 3.6 Mc.	The measured frequency response is ± 1 db from 10 cps to 7.5 Mc. See Figures 78, 79, 80, 81.

Table XI Table of Specifications (Continued)

Specification Item Number	Specifications	Test Results
3.5.2.10.3	Differential gain shall be less than 1.0 db.	Low Band: Unit 1 to 2, 0.7 db Unit 2 to 1, 0.2 db High Band: Unit 1 to 2, 0.6 db Unit 2 to 1, -0.1 db
3.5.2.10.4	Differential phase shall be less than ±2.0 degrees.	Low Band: Unit 1 to 2, 0.5° Unit 2 to 1, 0.5° High Band: Unit 1 to 2, 0.4° Unit 2 to 1, 0.75°
3.5.2.10.5	The total harmonic distortion over the system with input level of 1.5 volts shall be less than 2.5 percent.	Low Band: Unit 1 to 2, 0.6% Unit 2 to 1; 0.4% High Band: Unit 1 to 2, 0.9% Unit 2 to 1, 0.9% at RF level - 80 dbw and input level of 1.1 volts peak
3.5.2.10.6	The carrier-to-noise ratio of the system shall not be less than 40 db with a 60-db path loss inserted.	

$$\frac{C}{N}\,\text{db} = P_o\ \text{dbw} - 60\ \text{db} - [KTB + NF]\ \text{db}$$

Table XI Table of Specifications (Continued)

Specification Item Number	Specifications	Test Results

Low Band:

Radio Unit 1 to Radio Unit 2

$$\frac{C}{N}\,db = -1.8\ dbw - 60\ db - [-144 + 10 \log_{10} 20 + 10.5]\ db = 58.7\ db$$

Radio Unit 2 to Radio Unit 1

$$\frac{C}{N}\,db = 0\ dbw - 60\ db - [-144 + 10 \log_{10} 20 + 9.5]\ db = 61.5\ db$$

High Band:

Radio Unit 1 to Radio Unit 2

$$\frac{C}{N}\,db = -8\ dbw - 60\ db - [-144 + 10 \log_{10} 20 + 12.3]\ db = 50.7\ db$$

Radio Unit 2 to Radio Unit 1

$$\frac{C}{N}\,db = -5.5\ dbw - 60\ db - [-144 + 13 + 14.3]\ db = 51.2\ db$$

For receiver noise characteristics, see Figures 82, 83, 84, and 85

Power Consumption

108 volts rms input - 400 watts at 0.8 P.F.
120 volts rms input -- 460 watts at 0.8 P.F.
132 volts rms input - 520 watts at 0.8 P.F.

Transmitter Klystron Stability

Low-band klystron stability with line voltage variation of ± 10% is better than ± 0.002%. High-band klystron stability not measured.

Table XI Table of Specifications (Continued)

Specification Item Number	Specifications	Test Results
	Order Wire	
	Order-wire communication was tested and found to be of reasonable quality for limited use.	
	Antenna Pattern – See Figure 103.	

transformer through fuse F1, power switch S-1, interlock switches S-2, S-3, and thermostat switch S-5. The thermostat switch is normally closed; but should the temperature in the power supply rise as a result of overload or failure, the thermostat would shut off the power. Two neon power indicating lamps and a transient suppressor diode (CR54) are connected across the line. Switch S-4 is used to obtain the same transformer output voltages at either 50 and 60 or 400 cps. Each of the secondary windings is connected to a rectifier bridge and a capacitor filter.

b. Electronic Filter

The output of rectifier bridge No. 1 is applied through a fuse to the electronic filter section. In the electronic filter, resistors R2 and R3, and capacitor C2 provide a low ripple voltage which biases transistor Q3. The emitter current of Q3 biases the power transistors Q1 and Q2 connected in parallel. Since the base voltage of Q3 has essentially no ripple, the emitter voltage of Q3, and consecutively of Q1 and Q2, also contains very low ripple.

Thus, the output of the circuit is a low-ripple d-c voltage with most of the ripple components appearing across transistors Q1 and Q2. Variable resistor R3 provides the proper d-c operating point.

When the power is first applied, capacitor C1 is charged rapidly to its normal voltage. To prevent overloading transistors Q1 and Q2 when the capacitor C2 is being charged through R2 and R3, means must be provided to limit the maximum voltage across collector to emitter of Q1 and Q2. This limiting is performed by resistor R40 and zener diode CR5. When power is first applied, diode CR5 conducts until capacitor C2 becomes partially charged; then the diode is cut off and normal operation begins.

c. High-Voltage Supply Regulator

The output of the electronic filter is applied to a series regulator which drives the high-voltage inverter, and to the series regulator which supplies the 48 and 24 volts.

The series regulator for the high-voltage output consists of series regulating transistors Q4, Q5, and Q6, regulator driver transistors Q7 and Q8, and the reference and error amplifier circuitry including transistors Q9, Q10, and Q11.

The regulator output is adjusted by variable resistor R16. When the output is at exactly the proper voltage, the circuit is in equilibrium; when the output tends to deviate from this value the error is sensed, amplified, and applied to the regulating transistors in proper polarity to decrease the output error. This action may be traced easily in the circuit. When the equilibrium is established, all of the circuit transistors are biased precisely to give the proper output voltage, and the circuit is in a static state. When, because of a line voltage change, the output tends to increase, transistor Q11 conducts somewhat harder. The emitters of Q10 and Q11 become more positive. Since the base voltage of Q10 is fixed by the zener diode, transistor Q10 conducts less, and its collector becomes more positive, driving the base of Q9 in a positive direction. Transistor Q9 conducts harder and its collector voltage decreases. Consequently, the bias of regulator transistors Q4, Q5, and Q6 is lowered, resulting in a decrease of the output voltage. The output voltage thus remains essentially constant, with excellent regulation achieved because of the high gain of the control circuit.

d. High-Voltage Inverter

The output of the regulator is applied to the inverter, through inductors L1 and L2. When the power is first applied, resistors R18 and R19 provide initial biasing to transistors Q12 and Q13. When current begins to flow through either of the transistors (one transistor conducts somewhat better than the other, since no two transistors are identical), voltage is induced in the biasing winding (terminals 4, 5, 6), which biases the transistors in such a way as to increase the current and saturate the transistor which conducted first and cut off the other. The transformer is designed so that the core is saturated when the transistor saturates. Upon core saturation, the flux can no longer increase; induced voltage drops; and the conducting transistor begins to cut off. With the decrease of current, the polarity induced in the biasing winding is reversed, cutting off the previously conducting transistor and placing the second transistor in saturation. This process continues and the output becomes a square wave whose frequency is determined by the point of core saturation and by the applied voltage. There are two additional windings in the transformer, the outputs of which are rectified, filtered, and applied to the load. The purpose of inductors L1 and L2 is to prevent voltage spikes, which are generated during the transistor switching, from being fed back through the regulating circuits to the other outputs. Capacitors C6 and C7 prevent these spikes from damaging the transistors.

The output of winding 7-8, rectified and filtered, provides −730 volts with respect to ground. This value is set by adjustment of resistor R16.

The output of the winding 9-10 is also rectified and filtered, and provides 500 volts dc. The positive side of this output is connected to the -730-volt terminal so that -1230 volts, with respect to ground, is obtained. Two voltage adjusting networks are connected across the 500-volt output to give two independently adjustable outputs, each having a coarse and a fine control.

e. 24-48 Volt Supply

The output of the electronic filter is also connected to the 48-volt series regulator circuit. Q15 is the regulating transistor, Q16 is the driver, and Q17, together with the associated resistors and zener diodes, is the reference-setting and error-amplifier circuit. The voltage across zener diode CR34 is used as a reference and also as the 24-volt output.

f. Short-Circuit Protection

Short circuit protection, which cuts off the regulating transistor Q15 whenever a shorting condition exists, is provided to the 48-volt regulating circuit. In normal operation the voltage at the collector of Q14 is only about 1.5 volts above its emitter (emitter is at 48 volts). Capacitor C15 is charged through resistor R54 to approximately 43 volts, which is determined by zener diode CR31. The capacitor is connected to the base of Q14 through resistor R55 and diode CR32. Normally the transistor is in the cutoff condition and has no effect on the regulating circuit. However, when a short occurs, the output voltage drops, causing Q14 to conduct. With Q14 conducting, the base voltage of Q16 decreases, lowering further the output voltage. This process continues until Q14 is saturated and Q15 is cut off. Once the short circuit protection has been triggered, it will keep the 48-volt output shut off. To return the regulator to normal operation (after the short has been removed), the power must be shut off (permitting capacitor C15 to discharge), and then reapplied.

g. Filament and 90-Volt Supplies

Rectifier bridge No. 2 is connected to a regulating circuit to provide the 90-volt output and rectifier bridge No. 4, with its regulators, provides the 6.3-volt filament outputs. The regulator circuits, although somewhat different in appearance, are essentially the same in operation as described above.

h. 400 CPS Inverter

The filtered output of rectifier bridge No. 3 drives the 400-cps inverter. Operation of this inverter is identical to the 1-kc inverter described above. One secondary winding of the inverter provides 117-volt, 400-cps square waves to drive two blowers, one in the power supply, one external. The other winding provides a 60-volt, peak-to-peak square wave, which when clipped and adjusted to proper level, is used as a reference signal in the radio unit.

3. Test Results

The following results were obtained with all the power supply outputs loaded, and with a line variation of ± 10 percent.

1.	24 ± 0.01 volts at 125 ma,	with spikes:	60 mv peak-to-peak
2.	48 ± 0.01 volts at 340 ma,	with spikes:	120 mv peak-to-peak
3.	90 ± 1.2 volts at 55 ma,	with spikes:	30 mv peak-to-peak
4.	6 ± 0.07 volts at 1.2 amp,	with spikes:	60 mv peak-to-peak
5.	6 ± 0.07 volts at 1.2 amp,	with spikes:	60 mv peak-to-peak
6.	-730 ± 0.3 volts at 150 ma,	with spikes:	80 mv peak-to-peak
7.	-500 ± 0.2 volts at 0 ma,	with spikes:	80 mv peak-to-peak

Line current = 4.8 amp. at 120 volts rms, 60 cps

The spikes were of a very short duration (several microseconds) with repetition rates of 400 cps and 1 kc. The ripple, in all cases, was so low as to be insignificant in comparison to the spikes.

Except for the spikes, which are the result of layout and wire cabling, the power supply performs well. No failures occurred during the tests, as a result of component overstresses in normal operation. The few failures that did occur were a direct result of abnormal conditions imposed by the test personnel.

C. TRANSMIT BASEBAND AMPLIFIER

1. Mechanical Description

The transmit baseband amplifier is constructed on an epoxy-glass component board, using point-to-point wiring. Transistor heat is dissipated by convection-type heat sinks, so the unit can operate through the full temperature range with good reliability. Electrical connections

are made through a permanently attached cable, with the exception of the input signal for which a coaxial connector is provided. All components are readily accessible for trouble shooting or replacement. The unit is attached to the radio case by mounting screws and may easily be removed for maintenance.

2. Functional Description

The transmit baseband amplifier amplifies an input baseband signal to a level sufficient to directly frequency modulate the transmit klystron. It has an adjustable gain and wide frequency response to facilitate its use with various signal sources. It also has good linearity, intermodulation, differential gain, and differential phase characteristics.

The circuit diagram of the transmit baseband amplifier is shown in Figure 87. The amplifier consists basically of three, two-stage amplifier sections connected in cascade. Each of the sections contains a negative feedback loop, which connects the collector of the second transistor to the emitter of the first. This feedback arrangement lowers the output impedance, while increasing the input impedance. The configuration of the sections is well suited for cascading, since no appreciable amount of interaction results between them.

This configuration was adopted for the final design after considerable investigation of different approaches. The main difficulty encountered in design was the simultaneous requirement of very flat frequency response, very good linearity, and very small differential gain and differential phase. Two of the requirements were relatively easy to obtain; however, when an attempt was made to improve performance at the third, one of the first two was degraded. The adoption of the above described approach made it possible to meet all requirements.

The baseband input signal is applied to the first amplifier section through the gain-adjust potentiometer. The signal, amplified by approximately 11 db in this section, is coupled to the second section through a series peaking coil. The order-wire carrier is also applied to the input of the first section; however, because of isolating resistor R3, its amplification through the circuit is virtually independent of the potentiometer setting. The second section further amplifies the signal by an additional 12 db, and its output is coupled to the last section. The first two sections are powered by the 48-volt supply, which is common to several radio circuits. Capacitors C20 and C22, and resistor R49 provide isolation from these circuits.

The third section amplifies the signal by approximately 8 db. The maximum output is 26 volts peak to peak. To obtain this output level, a 90-volt power source is used. Resistor R50 and capacitor C21 serve to remove any stray signals picked up on the power line, as well as to lower the apparent power-supply impedance at high frequency (lead inductance).

The klystron reflector d-c voltage source is connected to the output of the amplifier through resistor R46. Clamping diode CR1, connected between the output of the amplifier (klystron reflector) and the -730-volt klystron cathode supply, provides protection for the klystron in case the reflector supply voltage is lost. The diode is normally reverse-biased, but with the loss of the reflector supply the diode starts to conduct, clamping the reflector at the cathode voltage. This action prevents the klystron from being damaged.

During the design of the amplifier, special emphasis was made to ensure proper operation with random selection of all components.

3. Design Goals

The following design goals were set for the transmit baseband amplifier:

Input Impedance	75 ohms ± 5%
Output Impedance	Unspecified
Output Load	50k resistive, paralleled by 25 $\mu\mu$f
Voltage Gain	Adjustable, 11 to 31 db
Dynamic Range	21 volts peak-to-peak min.
Frequency Response	±0.1 db from 60 cps to 5 Mc
	$^{+0.3}_{-1.0}$ db from 20 cps to 7.5 Mc referenced to 3.6 Mc

Intermodulation Distortion	53.5 db noise-to-noise ratio*
Differential Gain	Less than 0.2 db **
Differential Phase	Less than 0.2° **
Operating Temperature	-54° C to + 75° C

* Using a random noise signal covering 60 kc to 1200 kc at 2.35 volts rms at the amplifier output, pre-emphasized in accordance with CCIR Recommendation No. 275

** Measured with Kelly Test Set with amplifier output level of 6.25 volts peak to peak at 15 kc, and 3.2 volts peak to peak at 3.58 Mc.

4. Test Results

The following results were obtained in the testing of one of the final units:

Input Impedance	83 ohms
Output Load	50k resistive paralleled by 25 $\mu\mu$f
Voltage Gain	Adjustable to 31 db
Dynamic Range	26 volts peak to peak
Frequency Response	±0.1 db from 60 cps to 5 Mc
	$^{+0.4}_{-1.0}$ db from 20 cps to 9.5 Mc referenced to 3.6 Mc
Intermodulation Distortion	62 db noise-to-noise ratio
Differential Gain	0.1 db
Differential Phase	0.15°
48-volt d-c current drain	67 ma
90-volt d-c current drain	35 ma

The frequency response curve is shown on Figure 88. The small peaking at low frequency can be eliminated by minor changes in capacitor values. In general, this design has well met the established goals.

D. I-F PREAMPLIFIER

1. Physical Description

The AN/TRC-56 preamplifier is made up of three stages of 2N1742 germanium transistors. The complete unit is packaged in a 4-1/2" by 1-1/2" by 1-3/8" chassis and is intended to mount directly to the AN/TRC-56 Receiver-Mixer leg by plugging the unit into the mixer crystal contact. Sturdy mounting is accomplished by fastening the unit to the wave-guide with 10 screws. Two covers are provided and can be removed for the purpose of trouble shooting without disconnecting the preamplifier from the mixer leg. Power is supplied by a 3-wire cable. One lead is for ground, one is for +24 volts dc at approximately 7 ma, and the other is for monitoring the mixer crystal current. The i-f signal input is obtained by means of a teflon tip-jack and makes contact to the crystal i-f output when the unit is plugged to the mixer leg. I-F output from the preamplifier is provided by means of an RG-55/U cable permanently attached. An i-f test jack (J-1) is provided for the purpose of aligning the preamplifier or other associated i-f circuitry.

2. Design Goals

Certain requirements to be met were dictated by the specifications set forth in RADC 2562A. The specifications affecting the i-f preamplifier are the receiver noise figure, i-f bandwidth, and center frequency. So that a receiver noise figure of 10.5 db could be met, an i-f noise figure of approximately 3.5 db was selected. The temperature environment to be encountered in this equipment would dictate the desirability of silicon transistors. However, a survey was made and none could be found with a low enough noise figure to even warrant their use. In the final analysis, a Philco 2N1742 germanium transistor was decided upon. From this transistor, a noise figure of approximately 3.5 db could be anticipated. To overcome any problems which might be encountered in bandwidth and center-frequency variations, it was decided to take the approach of wideband amplifier stages followed by a bandpass filter circuit which would not be affected by temperature changes. In determining the design objective for the over-all preamplifier gain, it was decided to strive for a target gain of about 20 to 30 db. With this amount of gain, it was felt that the level would be sufficient to overcome any possibility of external i-f interference and

internal noise in early stages of the i-f amplifier, and provide sufficient isolation between the mixer and high-level stages of the i-f amplifier.

3. Functional Description

The schematic diagram of the AN/TRC-56 low-noise i-f pre-amplifier is shown in Figure 89. The preamplifier consists of the following:

a. A broadband, single-tuned network centered around 70 Mc, which serves as the i-f load for the mixer crystal and input network for the first stage of the preamplifier.

b. A low-noise transistor amplifier (Philco 2N1742) followed by a single-tuned configuration (approximately 20 Mc wide). Neutralization for this stage is achieved by coupling the signal at the collector back to the base by means of the bifilar-wound coil in the collector through the 0.75 pf coupling capacitor. The coil acts as a 1:1 transformer, and the coupling capacitor simulates the collector-to-base capacity which is approximately 0.75 pf. The bifilar coil output is properly phased so the returned neutralizing signal will cancel the internal collector-to-base feedback signal.

c. A second wideband i-f amplifier (also a 2N1742). In this amplifier, shunt resistive feedback exchanges gain for bandwidth, with a series inductance to peak the high-frequency response. The low-frequency response is dependent on the size of coupling and emitter bypass capacitors.

d. A cable driver stage which is also a wideband amplifier and employs a 2N1742. This is a relatively low-impedance output stage (about 100 ohms) which drives approximately 2 feet of 50-ohm cable terminated in 50 ohms. No adverse effects are experienced with this mismatch, since the cable length is relatively short.

4. Results

The breadboard model was designed and built with the approach described above and produced satisfactory results.

The measured noise figure of the initial breadboard was 4.5 db. This figure was obtained when a matching transformer was used to provide a match between the crystal i-f resistance and the optimum noise resistance of the preamplifier. It was later discovered that the measurement technique of the crystal and i-f resistances was in error. After correct measurements of crystal and i-f resistances of 290 and 130 ohms respectively, a noise figure of 3.7 db was obtained when both sources were matched.

However, it was found that some mismatch can be tolerated without seriously deteriorating the noise figure of the preamplifier. Hence, the preamplifier input network is a single-tuned network with no facility for tuning (since it is broadband due to the shunting effect of the crystal and transistor resistances) and results in a degraded noise figure of only 3.8 db. The breadboard preamplifier was tested in an r-f test setup in the range of 7.0 gc and the measured receiver noise figure was 10.25 db.

The over-all i-f bandwidth of the breadboard preamplifier was 20 Mc to the -3 db points, centered at 70 Mc. One control is provided for some adjustment of the bandpass, but its range is limited since it is in a low-Q tuned circuit. This control, however, has some desirable effects in adjusting for minimum differential phase of a color TV subcarrier without appreciably affecting bandpass characteristics. Over-all gain of the 3-stage preamplifier was 25 db with a ±2-db variation over the temperature range.

Four preamplifiers were built in the lab and one was subjected to the necessary tests. These tests were performed in a screen room with a typical system setup; i.e., with all r-f plumbing, klystrons (T_x and L.O.), i-f amplifier and video amplifiers, etc. Tests were made with the low- and high-band rf and all tests met specifications. The receiver noise figure with this unit measured 8.5 db in the low band and 11.2 db in the high band. This particular preamplifier, when tested in the first AN/TRC-56 radio unit, gave a noise figure of 9.5 db in the low band and 12.3 db in the high band. The second preamplifier had a receiver noise figure of 10.5 db in the low band and 14.3 db in the high band when incorporated with the second AN/TRC-56 radio unit. It is conceivable that the higher receiver noise figures obtained with the high band setup is due to noisy klystrons and higher conversion loss of the crystal mixer.

E. I-F AMPLIFIER

1. Physical Description

The i-f amplifier, when used in conjunction with the AN/TRC-56 i-f preamplifier, supplies the necessary demodulated video signal at 75 ohms impedance and contains circuitry to operate the AFC, carrier level monitor, and squelching. A special bandpass filter is incorporated as an integral part to achieve the desired i-f selectivity. All circuitry making up the i-f is a combination of printed cards and point-to-point wiring. A BNC-type jack is provided for the i-f input (J-1) and has a 50-ohm impedance. The output connector (J-2) is also a BNC-type jack. The amplifier

is housed in a 5-1/8" by 11-3/4" by 1-1/8" chassis. The printed-card stages are assembled into the chassis by sliding each one into its proper slot and then making the interconnections by means of point-to-point wiring. Controls which are brought out and made available at the top of the chassis are the squelch level adjust, carrier level indicator adjustment, and all controls necessary to adjust the i-f discriminator. Power to the unit is supplied by a 5-wire cable. One lead is for ground, one for + 24 volts dc at 100 ma, one for the AFC signal, one for monitoring squelch, and the other for monitoring the received carrier level.

2. Functional Description

A schematic diagram of the i-f amplifier is shown in Figure 90. The i-f amplifier consists of the following:

a. In the main i-f section, a first, second, and third i-f amplifier utilizing 2N1742's which are wideband amplifier stages. In these amplifiers shunt resistive feedback exchanges gain for bandwidth, with a series inductance to peak the high-frequency response. The low-frequency response of these amplifiers is dependent upon the size of coupling and emitter bypass capacitors.

b. A 20-Mc bandpass filter, consisting of lumped constant-k sections, which is preceded and followed by 150-ohm T-pads to isolate the impedance variations of the i-f amplifier.

c. Eight additional wideband i-f amplifiers as in a. above. The first seven employ 2N1742's, and the eighth employs a 2N834.

d. A discriminator driver stage which employs a 2N834 to supply the necessary gain and power for the discriminator circuit.

e. A discriminator circuit which has component parts on both a printed card and the i-f chassis. The variable components of the discriminator are those which are mounted to the chassis and are the Symmetry, Linearity, and Crossover controls.

f. A two-stage baseband amplifier section following the discriminator. The first stage, which is on its own printed card, uses a 2N706. This stage is a conventional class-A amplifier which has the negative

feedback signal inserted at its unbypassed emitter. The second or output stage, which is also on its own printed card, uses a 2N1613. The negative feedback signal is derived at its collector. The correct amount of feedback for obtaining proper over-all gain and output impedance is determined by the point at which this signal is derived. This is accomplished by a resistance divider network in the collector of the output stage. The output signal is capacitive coupled to output jack J-2, by paralleling two 120-μf capacitors.

g. Four stages of tuned i-f amplifiers employing 2N834's. The input to this amplifier chain is derived from the preamplifier output by paralleling it with the input of the main i-f amplifier section. These amplifier stages are also on individual printed cards. The stages employ double-tuned, transitionally coupled interstage networks with a resulting over-all bandwidth of approximately 20 Mc and over-all gain to the detector of about 45 db. Neutralization is used in every stage by paralleling the collector-to-base capacitance with a coil. Each interstage network contains a variable capacitor in the primary and secondary for aligning the bandpass.

h. A detector card consisting of two detector circuits. One detector is half-wave, employing the 1N914, whose rectified intermediate frequency is used to monitor received carrier level. The other detector is wired in a voltage-doubler fashion and utilizes two 1N914 diodes; its rectified i-f is applied to a 2N1613 emitter follower, also on the same card. The emitter-follower output dc triggers the squelching circuit.

i. A card containing the squelch circuit, which is a Schmitt trigger utilizing two 2N706 transistors. The output voltage of the squelch circuit is applied to the bases of the last two i-f amplifiers in the main i-f section.

3. Design Goals

Because of the wide temperature environment, it was decided that the best approach to the design of the i-f stages would be the wideband amplifier approach as used in the i-f preamplifier. To achieve the desired i-f bandpass characteristics, a bandpass filter was incorporated. Specifications regarding the filter were set as follows:

1. Center Frequency - 70 mc nominal

2. Input and Output Impedance - 150 ohms nominal

3. Ripple	-	0.5 db max. from 64 Mc to 76 Mc
4. Insertion Loss	-	-2 db max. from 64 Mc to 76 Mc
5. Bandwidth	-	Reference: 70 Mc Response: - 3 db max. at 60 and 80 Mc -20 db min. at 55 and 85 Mc -60 db min. at 45 and 95 Mc
6. Phase Linearity	-	0.4 deg. per Mc for 70 Mc to ±5 Mc.

The over-all gain necessary to achieve limiting in the last i-f stage would be from 90 to 100 db. It was decided to utilize 10 wideband stages with 2N1742's, one wideband stage with a 2N834, and a limiter-discriminator driver stage with a 2N834. The last two stages were required to handle and deliver power; therefore, the choice of the silicon 2N834 was made. Since a high degree of linearity was desired, it was decided to utilize a phase discriminator circuit with adjustable coupling and symmetry in addition to crossover frequency control. A final goal of 100 millivolts per megacycle for the discriminator slope was set. It was intended, in addition to supplying the baseband information, that the discriminator would also supply the d-c error signal for the AFC. For the receiver to be inoperative below threshold, a squelching circuit would be required, so it was decided that the last two i-f stages would be squelched. The circuit performing this function would be a Schmitt trigger, whose output would be applied to the bases of the last two i-f stages. So that the squelching circuit would not be triggered by any extraneous signals or noise, it was necessary that the detector circuit performing this function have its signal derived after the bandpass filter and before the point of limiting in the i-f amplifier, preferably between the fourth and fifth i-f stages. The bandpass filter was to be inserted between the third and fourth i-f stages. To provide the 75-ohm output impedance and baseband response required, it was intended to incorporate an emitter follower having a gain of unity after the i-f discriminator.

4. Results

Many problems were encountered in the initial design of the breadboard i-f amplifier. One of the first problems to be attacked was the baseband dropout section which began as a two-stage emitter follower

with built-in feedback. After making tests with this section, it was discovered that it had a high noise figure and improper output impedance. The final design of this section resulted in a two-stage amplifier with overall negative feedback. The output impedance was adjusted to 75 ohms and the gain to 1.2:1, with a frequency response of 60 cycles to 8 Mc at the −0.5-db points. This dropout section consists of one 2N706 followed by a 2N1613. Both input and output are a-c coupled with the output stage, which is collector loaded. The following are some of the pertinent test data:

Gain	-	1.2:1
Overload	-	2.5 vP-P input signal
Differential Phase	-	Approx. 0.125 degrees
Differential Gain	-	Approx. 0.01 db
Response	-	60 cps to 8 Mc minus 0.5 db
		20 cps (-3 db), 10 Mc (-1.0 db)

The following were the noise power ratio results with a 0.1-volt rms input level:

Measurement Frequency (kc)	NPR (db)
70	66.3
290	65.5
534	68.0
1002	69.0

The design of the i-f bandpass filter was not attempted in the laboratory. Instead, outside filter suppliers were solicited. The i-f amplifier design proceeded without this filter. Design of the intermediate frequency was performed according to the original approach without use of the bandpass filter, and tests revealed several problem areas. It was discovered that, while a discriminator slope over 90 mv per megacycle could be achieved, the two last 2N834 stages would be over-dissipated at the upper temperature limit as a result. Therefore, their operating levels had to be lowered

and the discriminator slope decreased to about 65 mv per megacycle. The squelching circuit originally planned worked quite well except for stability with temperature. It was stabilized by the use of a temperature-sensitive diode in the B+ supply and a sensistor in the emitter circuit of the Schmitt trigger. The signal levels required to trigger the squelching circuit were a difficult problem to cope with. It was originally intended to insert a detector circuit just after the bandpass filter which could be used for the trigger and received carrier level monitor. It was discovered that signals appearing at any of the 2N1742 i-f stages were either very low in level or had a very small dynamic range. This was a result of the low-level overload or saturation characteristics of the 2N1742. It was decided, then, to branch off into another i-f section which would employ the silicon 2N834 and tuned interstage elements. This was accomplished by taking part of the i-f input at J-1 (preamplifier output), amplifying the signal through four stages with a bandpass of approximately 20 Mc, detecting this signal, and applying it to the Schmitt trigger by means of an emitter follower. At this i-f output, another detector circuit was also employed to supply the necessary voltage for monitoring received carrier level. Curves of meter indication as a function of received carrier level are shown in Figures 91 and 92.

These circuits also required temperature stabilization. Complete tests of the i-f amplifier were performed when the prototype bandpass filter arrived. The supplier of the i-f filter could not comply with one specification. Instead of phase linearity of 0.4 degree per megacycle, the prototype produced 1.5 degrees per megacycle. The final test results, utilizing an elaborate bench mock-up to simulate a system and including such items as transmit baseband, all r-f components, preamplifier, i-f amplifier, and baseband dropout along with pre- and de-emphasis networks, are as follows:

1. Response - 20 cps to 7.5 Mc (±0.5 db)

2. Differential Phase - 0.8 degree

3. Differential Gain - 0.12 db

4. Noise power ratio and noise-to-idle-noise ratio with pre- and de-emphasis. (Signal at transmit klystron is 2.35 volts rms at reflector to simulate 240-channel FDM loading.)

Measurement Frequency (kc)	Noise Power Ratio (db)	Noise-to-Idle-Noise Ratio (db)
70	44	63
290	38	61
534	36	57
1002	41	57

The above test results indicate that no detrimental effects were experienced from the bandpass filter. In order that the filter function properly with this intermediate frequency, a pad had to be inserted at the input and output ends of the filter. This was necessary because the intermediate frequency did not present a constant impedance across the i-f band.

F. ORDER WIRE

1. Mechanical Description

The order-wire circuit is constructed on an epoxy-glass board using point-to-point wiring. The input and output phone jacks, and the push-to-talk and push-to-signal switches are mounted on a bracket to provide a single complete unit. When mounted in position in the radio case, the jacks and switches are accessible for operation through holes in the control panel. Connections to the radio are made with permanently connected laced wires which are tied to one of the radio terminal blocks. To remove the unit from the radio for maintenance, several screws are removed. This unit is also sprayed for fungus and humidity resistance.

2. Functional Description

The order wire provides voice communication between two radio units constituting a hop. In the order-wire system the audio signal amplitude modulates a carrier, the frequency of which is above the highest frequency component of the baseband signal. This modulated carrier is inserted into the baseband channel and transmitted. On the receiving side, the order wire, tuned to the carrier frequency, amplifies the modulated carrier and detects the modulation, providing the audio output. Signaling also is incorporated to alert the attendant that a call is being made.

The circuit diagram of the order wire is shown in Figure 93. When switch S1 is depressed, power is applied to the transmitter part of the order wire. The signaling oscillator generates a 10-kc signal which is applied to the base of the 8-Mc carrier oscillator transistor, causing AM modulation. The output of the carrier oscillator is coupled through resistor R19 to prevent oscillator loading.

When switch S-2 is depressed, power is applied to the audio-amplifier stage and to the carrier oscillator. A sound-powered phone connected to the input telephone jack provides the audio input signal. This signal is amplified by transistor Q2 and used to modulate the carrier. The values of resistors R8 and R12, and capacitors C5 and C7 are chosen to limit the audio response approximately from 200 cps to 3 kc.

On the receiver side the carrier is passed through a series-tuned circuit and applied to a two-stage, tuned-carrier amplifier. The series-tuned circuit prevents loading of the baseband signal and also helps to reject the noise from radio receiver. After amplification in the carrier amplifier the signal is demodulated. The audio signal is passed through the emitter-follower stage (Q54) and applied to the output. Inductance L54 and capacitor C58, connected to the emitter of Q54, are tuned to the 10-kc signaling frequency. This circuit has no effect upon the audio signal; however, when the signaling tone is received the emitter resistance appears bypassed and appreciable gain is realized at the collector of Q54. The signaling tone is then rectified and used to drive the buzzer-controlling transistor, Q53. The response of the order wire is limited to the normal telephone band (that is, 300 cps to 3 kc) to improve the signal-to-noise ratio.

The order wire may also be monitored and controlled from a remote position. Six wires connected to the remote positions are required, and permit a complete parallel operation.

3. Results

Tests of the order wire in the system during the final laboratory and field tests showed the performance of the order wire to be marginal. The output signal was noisy at times, but intelligible, and the order wire performed its function in the field.

G. RECEIVE BASEBAND AMPLIFIER

1. Mechanical Description

The receive baseband amplifier is constructed on an epoxy-glass component board with point-to-point wiring. Two connectors are mounted on brackets for the baseband input and output, and two pigtail wires, permanently attached to the board, connect to the power source. Two convection-type transistor heat sinks are used for heat dissipation.

2. Functional Description

The receive baseband amplifier is a wideband amplifier, designed to deliver up to 4 volts peak to peak across a 75-ohm load. It has excellent distortion, intermodulation, differential gain, and differential phase characteristics.

The circuit diagram of the receive baseband amplifier is shown in Figure 94. The input signal is applied to the base of transistor Q1 through the gain-adjusting potentiometer R1. Since the input impedance into the first amplifier stage is high, shunting resistor R2, together with the potentiometer, is used to establish the input impedance into the circuit at 75 ohms. The output of transistor Q1 is coupled into the second amplifier stage through a series peaking coil, L1. The third and output stage consists of two transistors connected in parallel. The bases and collectors of Q3 and Q4 are connected together, and the emitters are connected through resistors R19, R22 and the common resistor R23 to ground. Resistors R19 and R22 are bypassed by capacitors C7 and C8, and are used to provide stability of the quiescent operating points. There are two negative feedback loops in the circuit. One is from the collector of the output stage to the emitter of Q2, and the other from the emitter of the output stage to the emitter of Q1. The emitter resistors of Q1 and Q2 are partially bypassed to permit feedback connections and to provide high open-loop gain. The first feedback loop (collectors of Q3 and Q4 to emitter Q2) decreases the output impedance of the amplifier, and at the same time increases the input impedance to Q2. The second loop increases the input impedance of Q1, while having only small effect on the output impedance of the output stage. The degree of feedback given the two loops is so chosen as to obtain the output impedance of 75 ohms. The negative feedback improves the amplifier linearity and extends the frequency response. To control the upper end of the frequency response, a variable capacitor, C10, is provided which, together with the resistor R24, forms a time constant, adjustable for best response. The values and tolerances of the biasing resistors, and the operating points of the transistors were also chosen to give optimum performance in linearity, differential gain, and differential phase.

The amplifier is powered by a 48-volt supply, which is used also to power other radio circuits. Resistors R25 and R26; and capacitors C12 and C13 are used to decouple the amplifier from other circuits on the power line.

3. Design Goals

The following design goals were established prior to the start of the circuit design efforts:

Input Impedance	-	75 ohms ± 5%
Output Impedance	-	75 ohms ± 5%
Output Load	-	75 ohms, resistive
Voltage Gain	-	0 to 20 db adjustable
Dynamic Range with 75-ohm load	-	0 to 2.0 volts peak
Frequency Response	-	Referenced to 3.6 Mc ± 0.1 db 60 cps to 5 Mc $^{+0.3}_{-1.0}$ db 20 cps to 7.5 Mc
Intermodulation Distortion	-	53.5-db noise-to-noise ratio*
Differential Gain	-	Less than 0.2 db**
Differential Phase	-	Less than 0.2°**
Operating Temperature	-	-54° C to +75° C

4. Results

The following data show the performance of the receive baseband amplifier. All of the design goals, except the output impedance, were met:

Input Impedance	-	75 ohms ±5%
Output Impedance	-	87 ohms

* Using a random noise signal covering 60 kc to 1200 kc at 0.2 volt rms at the amplifier output.

** Measured with Kelly Test Set, with amplifier output 1.0 volt peak to peak at 15 kc, and 0.1 volt peak to peak at 3.58 Mc.

Output Load	-	75 ohms
Voltage Gain	-	20.9 db adjustable over full range
Dynamic Range	-	0 to 2.5 volts peak
Frequency Response	-	Referenced to 3.6 Mc ± 0.1 db 60 cps to 5 Mc $^{+0.1}_{-1.0}$ db 10 cps to 10 Mc
Intermodulation Distortion	-	69 to 75 db noise-to-noise ratio*
Differential Gain	-	Less than 0.05 db**
Differential Phase	-	Less than 0.15°*
D-C Power Consumption	-	73 ma at 48 VDC

The high output impedance was due to measurement error during design. It can be easily corrected by changing of a resistance value in the feedback circuit.

* Using a random noise signal covering 60 kc to 1200 kc at 0.2 volt rms at the amplifier output.

** Measured with Kelly Test Set, with amplifier output 1.0 volt peak to peak at 15 kc, and 0.1 volt peak to peak at 3.58 Mc.

H. AUTOMATIC FREQUENCY CONTROL (AFC) AMPLIFIER

1. Mechanical Description

The Automatic Frequency Control amplifier is assembled on an epoxy-glass component board, with point-to-point wiring. The component board is enclosed by a perforated metal case, on which are mounted line filter components and a terminal block. The components are mounted on both sides of the board for better utilization of space. The unit is mounted by two guide pins and several screws, and is easily removable for maintenance.

2. Operational Description

The AFC amplifier is a high-gain, stable d-c amplifier, used in the receiver for local-oscillator frequency control. Its purpose is to amplify the d-c component of the discriminator output to a level sufficient to correct for any frequency drift of the transmitter and the local oscillator klystrons in such a way as to place the resultant i-f signal in the center of its band. Control is accomplished by applying the output of the AFC amplifier in series with the power supply to the repeller of the receiver klystron. The AFC amplifier closes a feedback loop which tends to maintain the d-c component of the discriminator at zero.

A circuit diagram of the AFC amplifier is shown in Figure 95. To simplify description, the circuit is divided into five operational blocks: the magnetic modulator and filter, the 400-cps oscillator, the error amplifier, the reference amplifier, and the phase detector and filter.

The output of the magnetic modulator is proportional to the product of the 400-cps reference signal and the d-c input signal. Since the 400-cps reference signal is of a constant amplitude, the output is a 400-cps wave proportional in amplitude to the d-c input. When the input polarity is changed, the phase of the output is reversed.

Output of the magnetic modulator is passed through a series-tuned L-C filter to eliminate the undesired signal harmonics generated in the modulator. A back-to-back diode set (CR1, CR2), connected from a junction point between C2 and L1, and ground, limits the 400-cps signal to prevent overloading of the error amplifier.

The error amplifier is a three-stage d-c-coupled feedback amplifier which amplifies the 400-cps signal to the proper level. The biasing network is such that the three stages are interlocked by d-c feedback to provide for good stability.

Output of the error amplifier is coupled to the phase detector through a step-up transformer with a turns ratio of 1:5.

The 400-cps oscillator is a Butler type. Transistors Q5 and Q6 are coupled at their emitters by a series L-C circuit, and the feedback loop is completed by the connection from R38 through C20 to the base of Q5. The frequency of this type oscillator is virtually independent of transistor parameters, and is determined almost exclusively by the tuned section. Variable resistor R38 is adjusted to provide a proper output level. The 400-cps

reference signal is passed through an emitter follower (Q7) to reduce loading, and then applied to the magnetic modulator and reference amplifier. Before the reference amplifier, there is a phase and gain adjusting circuit. The phasing circuit is adjusted so that the phase of the error amplifier output and the phase of the reference amplifier output coincide.

The reference amplifier has an identical circuit configuration to that of the error amplifier, except for a few resistor values which are different to establish a lower gain. The output of the reference amplifier is also coupled to the phase detector through a transformer.

The input to the phase detector from the reference amplifier is a constant-amplitude 400-cps signal. During half of the period, when the point connected to diodes CR3 and CR4 is positive, the diodes are reverse-biased and cut off. During that time, the error-amplifier output is positive and charges C10 through R19. During the second half of the period, the diodes are forward-biased by output of the reference amplifier and conducting, so that the output of the error amplifier is essentially shorted by the low diode resistance. During this period, no voltage appears at the output of the error amplifier. Thus, the output of the error amplifier appears as a half-wave, rectified signal, which is filtered to provide the d-c output. When the polarity of the input to the magnetic modulator is reversed, the two inputs to the phase detector are exactly 180 degrees out of phase, and the above process takes place except that polarity of the output is reversed.

To prevent coupling between the AFC amplifier and other circuits being powered by a common power supply, a filter is provided on the +48-volt power line.

The transfer characteristic of the AFC amplifier is shown in Figure 96. The gain of the circuit was adjusted to the maximum value, but may be decreased if required.

When the circuit was connected in the system, the controlling action was obtained; however, low-frequency oscillation was also present. Investigation showed that several R-C constants in the feedback loop were sufficiently close to each other to cause this condition. The time constants were modified so that the oscillations became insignificant.

I. PRE- AND DE-EMPHASIS

1. Mechanical Description

The pre- and de-emphasis circuit is constructed on an epoxy-glass component board, with point-to-point wiring. The board is mounted by two guide pins inserted into holes on the radio case, and two screws. There are four coaxial cables connected to the board which carry baseband signal, and one single wire carrying order-wire signal.

There are four circuits mounted on the board. These are the pre-emphasis and de-emphasis circuits for multiplex operation, and the pre-emphasis and de-emphasis circuits for TV signals. Terminals are provided so that the desired net of circuits may be selected by soldering strap wires.

2. Functional Description and Test Results

A schematic diagram of the pre- and de-emphasis circuits is shown in Figure 97. Both the pre- and de-emphasis circuits are driven from a 75-ohm source and terminated by a 75-ohm load. The multiplex pre-emphasis circuit consists of a series L-C circuit shunted by a 271.5-ohm resistor. The circuit is connected in series with the signal and tuned to 1.5 Mc. As the signal frequency is increased at low frequencies, the impedance is essentially constant; but as the resonant frequency is approached, the impedance decreases. Past resonance, the impedance increases approaching a constant value determined by the resistor. The response of this circuit is shown in Figure 98.

The multiplex de-emphasis circuit consists of a parallel-tuned circuit shunted by a 271.5-ohm resistor. This circuit is also connected in series with the signal and is tuned to 1.5 Mc. At low frequencies, the impedance of the circuit is small, increasing to the value of the resistor as the frequency is increased to resonance, and decreasing as the frequency is raised past resonance. The frequency response of the de-emphasis is shown on Figure 98. The frequency response of the pre-emphasis and de-emphasis circuits are so matched as to give a flat response when both of the circuits are used in a system.

The TV signal pre-emphasis and de-emphasis circuits are four terminal-shunted T-type circuits, with a constant input and output impedance of 75 ohms. The pre-emphasis circuit attenuates the low frequencies by 20 db more than the high frequencies, whereas the de-emphasis attenuates

the high frequencies more. The crossover frequency, the frequency at which the attenuation of both circuits is 10 db, is 1 Mc. Frequency response of the TV pre- and de-emphasis is shown in Figure 99.

The measured frequency response coincided with the theoretical. The circuits used are in accordance with the CCIR recommendations.

Attenuation for the multiplex pre-emphasis circuit is:

$$A_{mux} = 10 \log_{10} \left[1 + \frac{6.90}{1 + \frac{5.25}{\left(\frac{fr}{f} - \frac{f}{fr}\right)^2}} \right] \text{db}$$

where, fr = 1.5 Mc.

Attenuation for the TV pre-emphasis circuit is:

$$A_{t_v} = 10 \log_{10} \left[1 + \frac{99}{1 + 10 \left(\frac{fb}{f}\right)^2} \right] \text{db}$$

where, fb = 1.0 Mc.

J. METERING AND TEST FACILITY

1. Mechanical Description

The Metering and Test Facility assembly consists of an epoxy-glass component board, metal plate, and two rotary switches. The rotary switches and the component board are mounted on the metal plate to make this unit one complete assembly. The assembly is mounted flat to the inner plane of the control panel, with the switch knobs mounted on its face. The meter is mounted directly on the control panel and two short wires connect it to the circuit. The permanently connected cable connects the various voltages to be measured through the terminal blocks of the radio unit.

2. Operational Description

The circuit diagram of the Metering and Test Facility is shown in Figure 100. The circuit is designed to permit monitoring of all power-supply voltages, klystron reflector voltages, AFC amplifier output, received carrier level, mixer crystal bias, i-f squelch status, and output levels of both baseband amplifiers. The circuits in this assembly consist of series dropping resistors for d-c measurements, and rectifiers plus resistors for a-c measurements. Meter readings are normalized so that only one scale is used on the meter face.

K. R-F COMPONENTS

1. Preselectors

Each receiver is equipped with a two-section direct-coupled post-type waveguide preselector. The preselectors are noteworthy because their bandwidth remains essentially constant across the entire tuning range. The low-band preselector bandwidth typically changes 8 Mc when tuned from 7.125 to 8.4 gc. A conventional quarter-wave-coupled, post-type waveguide preselector would vary as much as 60 Mc in bandwidth across the same band. A new method of tuning, whereby the broad wall width of the waveguide is effectively varied by means of two nonresonant cavities, was used to achieve the constant bandwidth. Typical responses of the filters are shown in Figures 101 and 102.

2. Isolators

The transmit klystron in both the high and low band is protected from antenna mismatches by a single-magnet ferrite field-displacement isolator. The isolator used in the 7.125 to 8.4 gc band provides 30 db minimum reverse attenuation with a maximum loss in the forward direction of 1 db. The high-band (14.0 to 15.4 gc) isolator has a minimum reverse attenuation of 40 db and a maximum forward loss of 1 db.

3. Frequency and Power Monitor

The Frequency and Power Monitor permits checking of the frequency and power output of the transmit klystrons. A portion of the transmitted signal is coupled from the main waveguide by a directional coupler and passed toward a cavity tuned to the desired transmit frequency. If the signal frequency differs from that of the cavity, the signal is reflected

back toward a crystal detector and the transmit power is read on the meter. If the signal frequency is the same as that of the cavity, the signal is absorbed by the cavity and a minimum output is seen on the meter.

A vernier tuning adjusts the cavity frequency approximately ±10 Mc so that the difference between the transmit frequency and the desired frequency can be ascertained. The vernier also permits detuning the cavity so that transmit power can be read with the transmitter at the correct frequency.

During system tests it was determined that the cavity Q on the low-band unit was too low for good frequency indication. The high-band cavity exhibited what appeared to be multimoding resulting in erratic performance. An external horn and detector were then provided for external monitoring of the transmitter.

4. Klystrons

Two types of klystrons are used, the VA244D for the 7.125 to 8.4 gc band, and the VA92F for the 14.0 to 15.4 gc band. Both are direct-modulated reflex klystrons with waveguide output, and both use forced-air cooling. The VA244D employs external cavity tuning and the VA92F uses internal cavity tuning. The VA92F is subject to microphonic frequency modulation which degrades its TV transmission performance. However, this does not affect its multiplex transmission capability. Eight each of the klystrons were evaluated with the following results:

<u>VA244D</u>

P_o	- 1.0 to 1.3 watts at 7.125 gc 0.69 to 0.85 watt at 8.4 gc
Modulation Sensitivity	- 325 to 473 kc/volt
Mode Bandwidth	- 27 to 44 Mc
Cathode Current	- 74 to 75 ma at -730 volts
Reflector Voltage	- -260 to -485 volts with respect to cathode

<u>VA92F</u>

P_o	- 130 to 290 mw at 14.0 gc 164 to 291 mw at 15.4 gc
Modulation Sensitivity	- 650 kc/volt average
Mode Bandwidth	- 28 to 44 Mc
Cathode Current	- 27 to 39 ma at -750 volts
Reflector Voltage	- -290 to -477 volts with respect to cathode

5. Antenna

The Antenna is of a parabolic type, two feet in diameter, and is an integral part of the radio unit. It has a replaceable circular center section approximately one foot in diameter, on which the transmitter and receiver feeds are mounted. The two feeds are cross-polarized to provide 30-db isolation. A high- and a low-band feed assembly is provided with each antenna, permitting its use over either of the two bands.

The Antenna has a beam width of about three degrees, with the sidelobes over 20 db down. The radiation pattern measured under field conditions at the high-frequency band is shown in Figure 103.

6. Local-Oscillator Injector and Mixer Assembly

The receiver preselector is followed by a single-ended mixer which was chosen because of its simplicity. The very low noise content of the local-oscillator klystron used made a balanced mixer or local-oscillator filter unnecessary. Final testing, however, indicated that the inclusion of a local-oscillator filter would have resulted in much easier receiver klystron tuning.

The local oscillator is fed to the mixer through an adjustable attenuator and a directional coupler, permitting optimum adjustment of local-oscillator power for best receiver noise figure. A dielectric phase shifter between the preselector and the mixer crystal is used to properly phase the image signal (a mixer product due to the presence of both the local-oscillator and i-f signals in the mixer diode). This phasing adjustment eliminates mixer i-f output impedance-to-preamplifier input impedance mismatch, removing this source of noise figure degradation.

SECTION IV

RADIO-MULTIPLEX SYSTEM TESTS

A. IN-LAB TESTS

Both units, radio and multiplex, were connected and operated as a duplex system in the laboratory.

Two coaxial cables are used to provide the baseband signal in-and-out interconnection of the two units. Figure 104 shows the test setup. The units were separated a maximum distance of 20 feet during the system tests. The maximum distance at which the equipments could be placed from one another before noticeable degradation of performance occurred was not measured because of limited testing time. No problems should be encountered with greater distances between the two equipments since interconnection impedances have been standardized at 75 ohms. The two coaxial cables from the units were terminated with standard connectors so that a patch cable of any desirable length could be interspaced.

When both units are to be connected together, the level adjustments are made on the radio unit. The multiplex transmitter baseband output is adjusted for full on. The receiver and transmitter of the multiplex are fixed and the necessary level adjustments are then made on the radio. The radio was purposely designed with adjustable inputs and outputs to give it added flexibility.

The only degradation of performance when the multiplex was connected to the radio was in the signal-to-noise ratio. Table XII is a reading of the signal-to-noise ratios for each channel when the multiplex is connected to the radio. The noise contribution of the radio is due to two sources: high frequency signals on the B+ line from the power supply, and the inherent noise in the receiver front end. The exact separate contributions from these two sources could not be determined.

The harmonic distortion and crosstalk, listed in tables VII, VIII and IX, are readings taken with the multiplex and radio connected. It was found that the multiplex is capable of the best performance when there is 3 db of gain at baseband frequencies between the two units. Crosstalk and harmonic distortion are the two characteristics most affected. The multiplex connected in a back-to-back configuration had crosstalk and distortion

Table XII AN/TRC-56 Multiplex
Noise Measurement

REFERENCE:

Philco Record Book: 11729
Date: August 21, 1962

TEST CONDITIONS:

a. Multiplex and radio connected
b. 0 dbm input, -3 dbm output at 1 kc, channels 1 through 12
c. -16 dbm input, -3 dbm output at 1 kc, channels 22 through 24

TEST RESULTS:

	UNIT A TO UNIT B				UNIT B TO UNIT A			
	Noise		S/N Ratio		Noise		S/N Ratio	
Channel Number	F1A (dba*)	Flat (dbm)	F1A (db)	Flat (db)	F1A (dba*)	Flat (dbm)	F1A (db)	Flat (db)
1	33.0	-47.5	52.0	44.5	34.0	-47.0	51.0	44.0
2	41.5	-42.0	43.5	39.0	35.5	-45.5	49.5	42.5
3	35.7	-43.3	49.3	40.3	37.2	-46.8	47.8	43.8
4	41.0	-44.0	44.0	41.0	32.2	-47.8	52.8	44.8
5	34.7	-42.3	50.3	39.3	33.2	-41.8	51.8	38.8
6	40.6	-33.0	44.4	30.0	34.2	-36.3	50.8	33.3
7	36.7	-44.3	48.3	41.3	31.8	-51.7	53.2	48.7
8	41.2	-43.3	43.8	40.3	33.5	-49.5	51.5	46.5
9	38.4	-43.8	46.6	40.8	32.8	-47.7	52.2	44.7
10	40.2	-43.8	44.8	40.8	36.1	-42.4	48.9	39.4
11	42.7	-36.3	42.3	33.3	34.8	-42.7	50.2	39.7
12	44.7	-32.3	40.3	29.3	33.0	-39.0	48.0	36.0
22	35.5	-41.5	49.5	38.5	37.0	-45.0	52.0	42.0
24	34.0	-39.0	51.0	36.0	30.0	-45.0	55.0	42.0

*Referenced to 0 dbm transmission level.

increased by approximately 5 db in most channels over the values obtained when used with the radio. This is due to lower distortion in the transmit and receive envelope-detector at higher level inputs, and the crosstalking signal produced in the transmit and receive signaling driver with low-level inputs. It was found in the laboratory that poor ground connections between the equipments could increase the noise, hum, and ripple in each channel. An indication of the subjective quality of the channels was that, in many cases, the EE8 field phones were the determining factor in the sound of a channel.

B. FIELD TEST

The multiplex-radio system was field-tested after laboratory testing had been completed. The two locations were nearly two miles apart. One location was a Philco field site in Abington, Pa., and the other was the roof of a 12-story apartment house. The tripod of the radio unit at the field site was placed on the ground. The line-of-sight elevation of this unit was approximately four degrees, and passed close over the tops of several trees and buildings.

Order-wire communication between the two sites was established within five minutes after the equipments were placed and warmed up. The multiplex equipments were then connected and contact established.

The field-test results were very satisfactory. The radio units were detuned and retuned, an antenna pattern was taken (see Figure 103), and multiplex channel signal-to-noise ratios were measured. The performance of the equipments in the field was equivalent to that in the laboratory. No equipment failure occurred, except for a shorting of a mica insulator in the radio power supply which, however, did not cause any component failures. The insulator, damaged during initial construction, was replaced and communication was re-established.

The equipment was in the field two weeks, under extreme (for this area) conditions. Temperatures ranged from 91° F down to 40°F, with both dry and rainy periods.

After the field tests were completed, the equipments were rechecked in the laboratory before shipment.

SECTION V

CONCLUSIONS

Analysis of the design and testing programs on the AN/TRC-56 Radio Set resulted in the following conclusions.

A. RADIO

1. Power Supply

The power-supply layout is such that the accessibility to many of the components for servicing or replacement is difficult. Also, because of the layout and wire cabling, undesired voltage spikes appear at the outputs. A change in the layout would overcome the above difficulties; however, little weight decrease may be expected from this change. If, however, the specifications were modified to require the unit to operate from a 400-cps line only, considerable savings in size and weight could be realized.

In general, the electrical performance of the power supply (neglecting the spikes) was very satisfactory.

2. Pre-emphasis and De-emphasis Network

The electrical performance of this unit is good; however, its mounting position in the radio unit is not convenient. The size of this unit could be decreased by constructing it in a three-dimensional modular form. This modification would also permit its mounting in a more convenient place.

3. Transmit Baseband Amplifier

The electrical performance of the transmit baseband amplifier is good. Component layout is acceptable, but small modification could be beneficial. The mounting location is good and all components are readily accessible.

4. Frequency and Power Monitor

Neither the low- nor the high-band unit is satisfactory. Although power output can be monitored, the units are not adequate for transmitter frequency tuning.

5. Klystron

Both of the klystrons meet the tunable bandwidth requirements; however, they are the linearity limiting components in the system. The low-band klystron, although rated at one watt nominal, is not capable of one watt at the upper end of the band. The tuning screw has no mechanical stop; as a result, the klystron may be damaged during tuning by turning the screw too far. The high-band klystron exhibits microphonic sensitivity which is generally not desirable.

6. Isolator

Performance of the isolator is good in all respects.

7. Antenna

Performance of the antenna was found satisfactory.

8. Preselector

The performance of both the high-band and low-band preselectors is very good in all respects. Both units cover wide tuning ranges and have uniform bandwidths over the entire bands. The bandwidth variations over either of the two bands is of the order of ±8 percent.

9. Mixer-Preamplifier

Performance of the mixer-preamplifier unit, in general, is very good. The noise figure in the high-frequency band of operation is somewhat high, but perhaps may be improved by better crystal matching and local-oscillator filtering. L. O. filtering has not been provided, and its lack results in difficulty of tuning the L. O. to the proper operating point.

10. I-F Amplifier

Performance of the i-f amplifier, in general, is reasonably good; however, its discriminator slope is lower than anticipated. The frequency response tends to degrade at high r-f power levels due to overload in transistors. This effect and the number of transistors in the circuit could possibly be decreased if circuit modification could provide a wider dynamic range. The wider dynamic range would eliminate the need for the parallel stages used now to drive the metering and squelch circuits.

11. Receive Baseband Amplifier

Performance of the receive baseband amplifier is good. The somewhat high output impedance can be corrected without difficulty by a change of resistor value in a feedback loop. All other operating characteristics were good, and no other modifications are needed.

12. Automatic Frequency Control Amplifier

The performance of the AFC amplifier is very satisfactory in all respects. The unit can possibly be made smaller by reducing the number of components, including transistors, without degradation of performance.

13. Metering and Test Facility

The metering unit performed well in all respects.

14. Order Wire

The order-wire performance in the system, although adequate for its limited task, is not on a par with the rest of the system. Improvement of the performance would be desirable.

15. Temperature Control

The temperature-controlling air vent system on the radio unit, although never tested over the entire temperature range, is calculated to be insufficient to hold the transmitter frequency within specification limits over the -54°C to + 52°C ambient. If transmitter frequency stability is to be held over the entire temperature range to the specified degree, the temperature control system may have to be modified.

16. Weight and Size of the Radio Unit

There is little hope of considerable weight or size reduction of the radio unit unless some of the following changes are made:

a. The unit is restricted to operate from a 400-cps line only.

b. Requirement for the band interchangeability is eliminated.

c. The power supply is packaged separately.

d. The antenna is packaged separately.

The radio dimensions in the present configuration were essentially dictated by the antenna size.

B. MULTIPLEX

The experimental AN/TRC-56 multiplex, excluding weight, met the specifications with minor exceptions. The one real shortcoming of the equipment is its over-all weight. However, of all the characteristics of the unit, the weight factor could be changed the most. The major contributions to the over-all weight were the universal type of application for which it was designed in signaling and dialing, and the over-all packaging.

In field use of this equipment, it would not be likely that both the a-c ringdown signaling and E & M dialing circuitry be included in one package. Use of one or the other would reduce the bulk of the Signaling and Termination Assemblies by one half. Options for a-c ringdown operation or E & M dialing operation could still be provided by making their respective cards interchangeable. Separate packaging of the signaling and termination units would further reduce the over-all case size and weight.

Packaging of the developed unit provided for adding an additional 10 data channels. Eliminating this requirement would also reduce size and weight.

All these steps would increase the portability of the equipment by a great degree without deteriorating the performance of the unit.

SECTION VI

RECOMMENDATIONS

Based upon experience gained in the design phase and measurements made during the laboratory and field testing phases, the following recommendations are made for future equipments:

A. RADIO

1. The frequency and power monitor for the transmit klystron should be redesigned. The new design should include a broadband (20-Mc) indication which would permit easy initial coarse tuning of the klystron. In addition, it should contain a fine-tuning indication which would permit exact final tuning of the klystron.

2. A bandpass filter should be incorporated in the mixer arm attached to the local-oscillator klystron. This filter would permit tuning the local oscillator to within the receiver AFC capture range without requiring an incoming r-f signal. This feature would be invaluable during system installation when communications had not yet been established (antenna alignment not yet completed) and when a local-oscillator klystron is replaced. It would also prevent incorrect operation of the local oscillator 35 Mc(one-half intermediate frequency) away from the received signal with the resultant degradation in system noise performance. It would eliminate noise contributions from the local oscillator which may be a significant contributing factor in the high-band noise figure.

3. The radio power supply should be made an entity in itself, separate from that portion of the radio which contains the antenna. This recommendation is based on the size and weight involved. The horizontal and vertical dimensions of the present, and any future, radio unit are dictated by the antenna diameter and are therefore unchangeable. The front-to-back depth of the present unit is determined primarily by the power supply. If the power supply were removed, the depth and weight of the remaining unit could be greatly reduced. The consequent weight reduction would permit a similar weight reduction of the tripod. Making the power supply a separate unit would allow a much better form factor to be achieved for it. At present it is three-dimensionally wrapped around the antenna contour. Removal of the power supply from the case would also prevent direct radiation of power-supply generated spikes into the other circuitry.

4. Radio sets for operation either in the low band (7.125 - 8.4 gc) or the high band (14.0-15.4 gc) should be considered as separate equipments without the need for interchangeability. This would result in more efficient mechanical packaging of the radio, with a consequent weight reduction.

5. The transmitter frequency stability requirement should be revised to cover any 30 °C temperature range within the over-all operating ambient of -54 °C to + 52 °C. This would reduce the complexity of the klystron temperature control problem and would still be compatible with the intended usage of the equipment. A temperature change greater than 30° C is extremely unlikely to occur in any given locale over a reasonable length of time (maintenance interval). When the equipment is transported to a new locale, normal practice would be to check transmitter tuning upon installation at the new site.

6. Either the television transmission capability (implicit in the differential phase and differential gain specifications) should be deleted, or the requirement for simultaneous order wire and TV transmission should be deleted. The order wire could then be operated at voice frequencies on the radio baseband, yielding several distinct advantages. It would remove the order wire from its present 8-Mc baseband frequency, where the characteristic noise of an FM receiver is highest. It would eliminate the need for modulating the order-wire information onto a carrier (as is done now with a resulting 10-db degradation in S/N). Compatibility of order-wire characteristics with those of equipments such as the AN/TRC-66 would be easy to obtain. It would make party-line order-wire operation on a system possible without the present push-to-talk arrangement.

7. The transmitter power required into the antenna should be revised from the present one-watt design goal to 600 milliwatts for the low band (7.125 - 8.4 gc). The klystron now used is a state-of-the-art device because of its extremely wide tuning range (1300-Mc tuning range compared to a 300-Mc tuning range of klystrons in general usage). To achieve the present design goal would require using more than one tube (probably four) to cover the tuning range. The resulting logistics problem would negate one of the major operational features of the equipment.

8. For production equipments, the required receiver noise figure (including preselector loss) for the low band (7.125 - 8.4 gc) should be changed to 12 db. The required receiver noise figure for the high band (14.0-15.4 gc) should be changed to 14 db. The recommended values are based on adding approximately 1 db margin to the theoretically calculated values. The theoretical values (assuming resistive image termination) are given by:

$$N = L_p L_c \left[N_{if} L_m + N_r - 1 \right]$$

where:

N = noise factor of receiver

L_p = preselector filter loss

L_c = mixer diode conversion loss

N_{if} = i-f amplifier noise factor

*L_m = mismatch loss between mixer and i-f

N_r = output noise ratio of mixer diode

* A function of Z_{if} and Z_c, i-f input, and diode output impedances, respectively

The pertinent equipment characteristics are:

	LOW BAND	HIGH BAND
L_p	1.26:1 or 1.0 db	1.41:1 or 1.5 db
L_c	2.82:1 or 4.5 db	4.47:1 or 6.5 db
N_{if}	2.52:1 or 4.0 db	2.52:1 or 4.0 db
L_m	1.17:1 or 0.7 db	1.26:1 or 1.0 db
N_r	1.60:1	1.30:1
Z_{if}	400 ohms	450 ohms
Z_c	335 - 465 ohms	365 - 565 ohms
Diode	1N415E	1N78D

For the low band

$$N = 1.26\,(2.82)\left[(2.52)\,(1.17) + 1.6 - 1\right]$$

$$= 12.6, \text{ or } 11.0 \text{ db}$$

For the high band

$$N = (1.41)\,(4.47)\left[(2.52)\,(1.26) + 1.3 - 1\right]$$

$$= 21.9, \text{ or } 13.4 \text{ db}$$

The calculations as noted previously are based on resistive termination of the image frequency. The AN/TRC-56 receiver has provisions for reactively terminating the image. This feature has the effect of reducing the mixer-diode conversion loss. The amount of improvement achievable is dependent upon the broadband conversion loss of the diode, increasing as the conversion loss decreases. For the diodes used, the improvement is 0.5 db in the low band and 0.2 db in the high band.

9. The 3-db down point at the high-frequency end of the over-all one-hop frequency response should be set at 6 Mc, with 100 kc taken as the reference. The response would still be almost six times that required to carry the multiplex signal, and adequate to carry typical TV with either intercarrier sound or a separate sound subcarrier. This would yield a reproducible production equipment.

B. MULTIPLEX

1. If no more than two data channels are required for most applications, the expandability requirement should be deleted. This would decrease the size of the multiplex package and, as a consequence, decrease its weight.

2. The Signaling and Termination units now provided with each voice channel should be packaged in a separate case, use of which would be dependent upon whether or not signaling is required. If the Signaling and Termination units cannot be put in a separate case then separate units for a-c ringdown and for E & M dialing should be supplied on an interchangeable basis. These recommendations are made with a view toward making the multiplex more easily portable from a size and weight standpoint.

3. The crosstalk specification should be revised to specify a method of measurement, a crosstalk design goal, maximum crosstalk, and a maximum number of channels which can exceed the design goal but be less than the specified maximum crosstalk. The value of the recommendation can be seen by examining the tables of crosstalk measurements presented in an earlier section of this report.

4. A receiver baseband amplifier with a large dynamic range should be provided. This would raise the per-channel input to the Transmit-Receive channel cards on a common amplification basis, eliminating an equivalent per-card gain requirement. This would remove as much as three stages of amplification on each Transmit-Receive card.

5. A more efficient method of combining channels at the baseband frequency output of the channel transmitters should be devised. The present low levels at this point make the channels susceptible to pickup of standard AM broadcast stations.

C. SYSTEM

The system channel noise specification should be revised in a manner similar to that recommended for the crosstalk measurement.

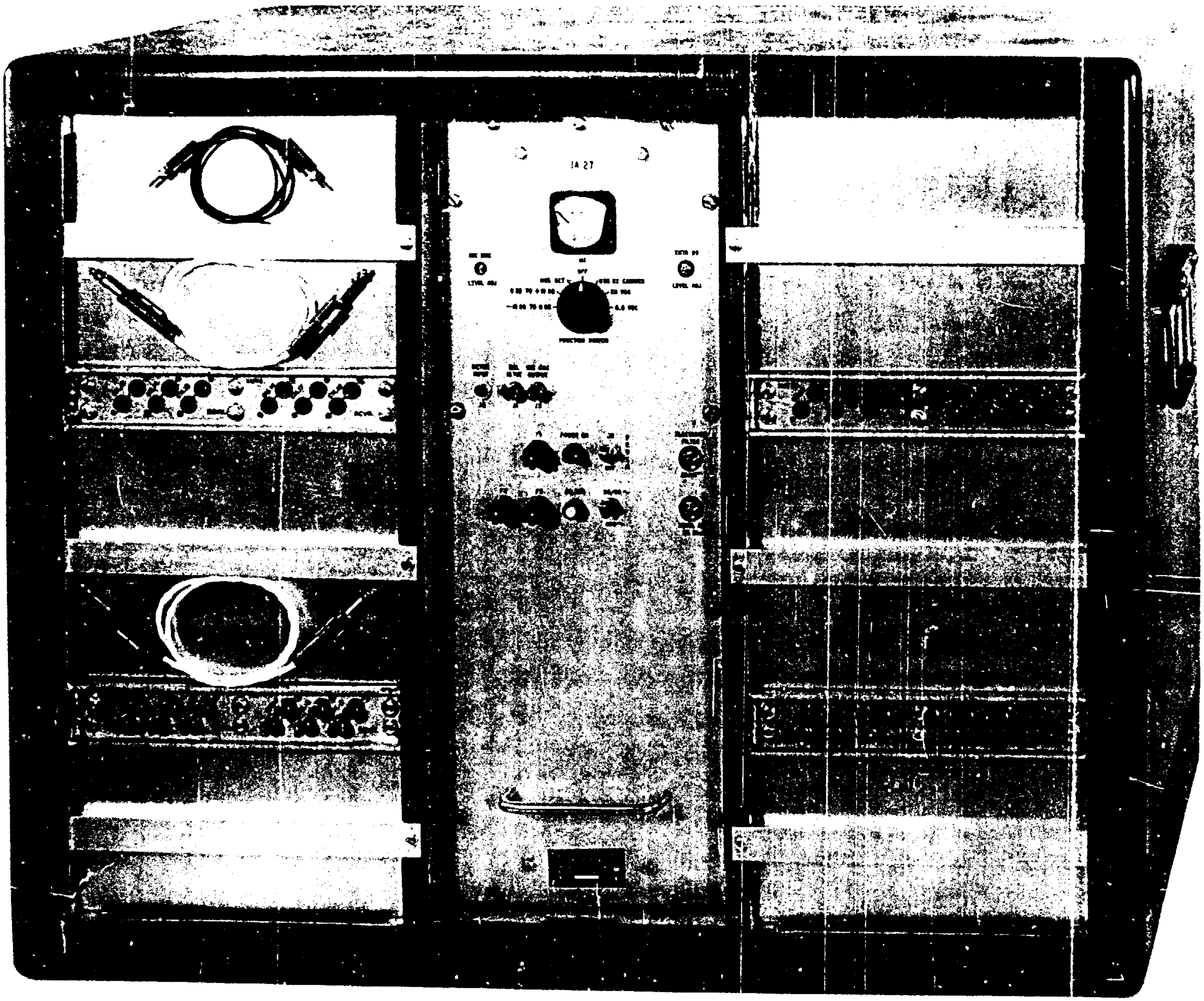
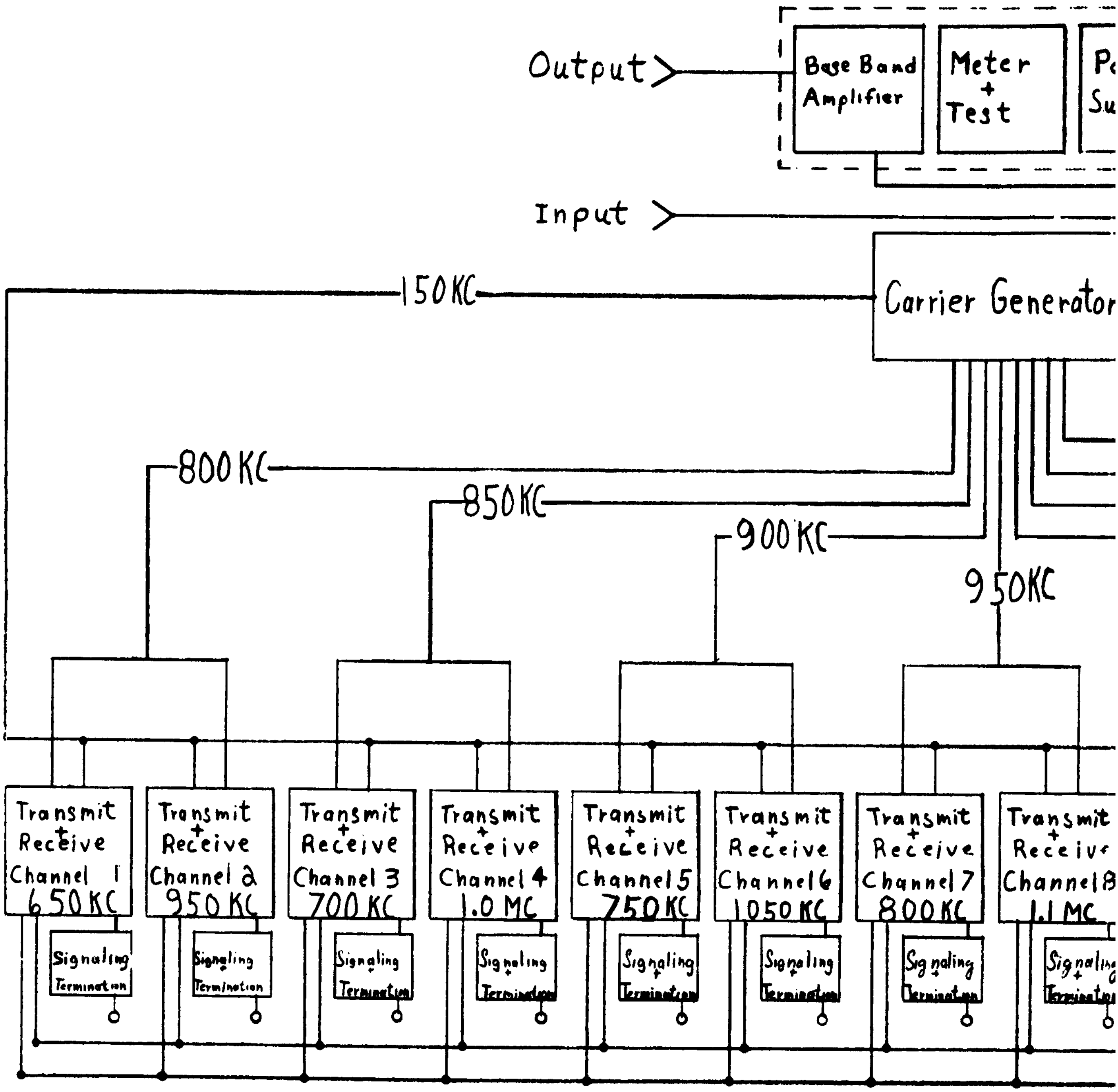


Figure 1 Multiplex Unit, Front View, Cover Removed


Output
Base Band Amplifier
Meter + Test
Input
Carrier Generator
150KC
800KC
850KC
900KC
950KC
Transmit + Receive Channel 1 650KC
Transmit + Receive Channel 2 950KC
Transmit + Receive Channel 3 700KC
Transmit + Receive Channel 4 1.0MC
Transmit + Receive Channel 5 750KC
Transmit + Receive Channel 6 1050KC
Transmit + Receive Channel 7 800KC
Signaling + Termination

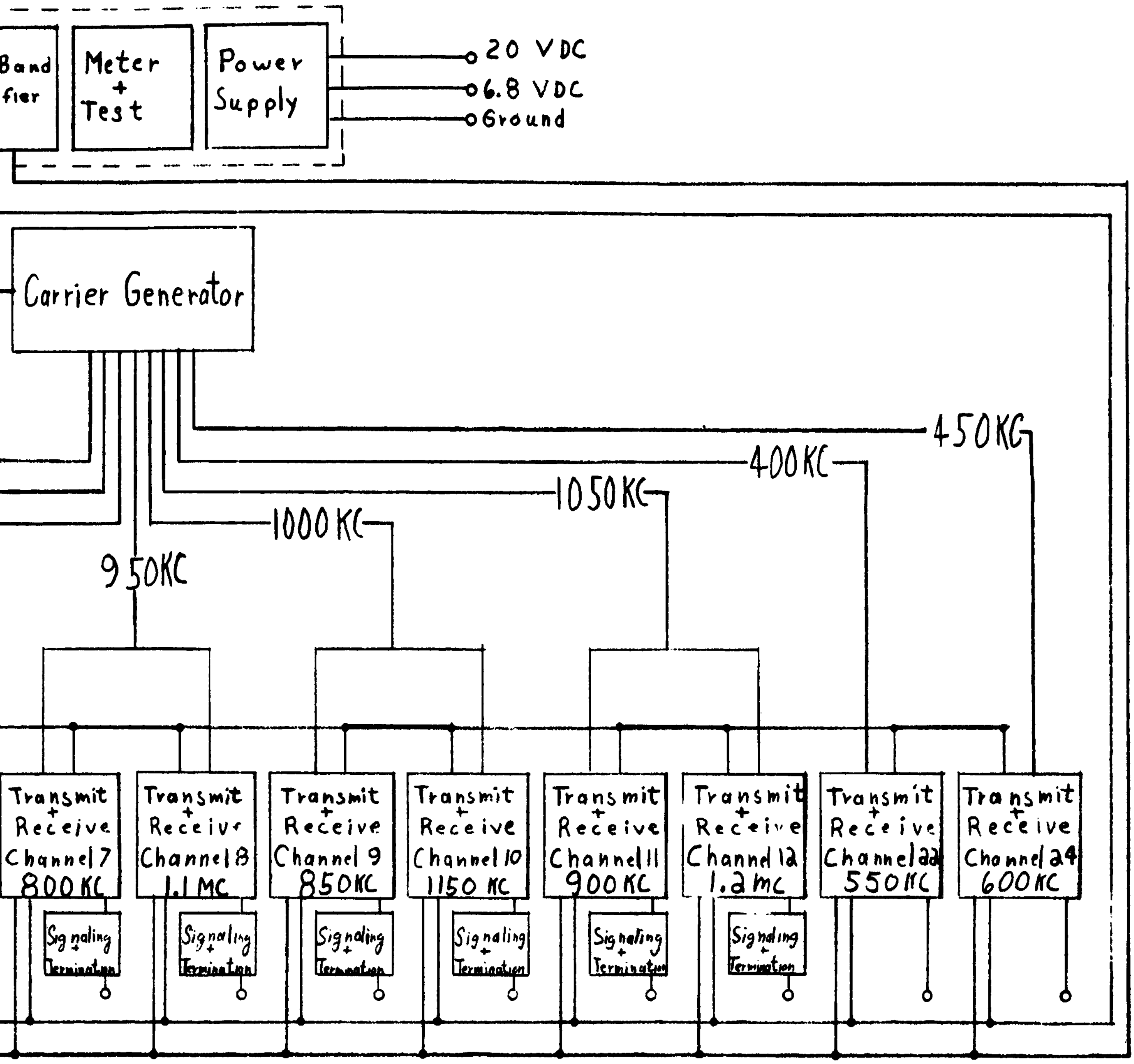


Figure 2 Simplified Multiplex Interconnection Diagram

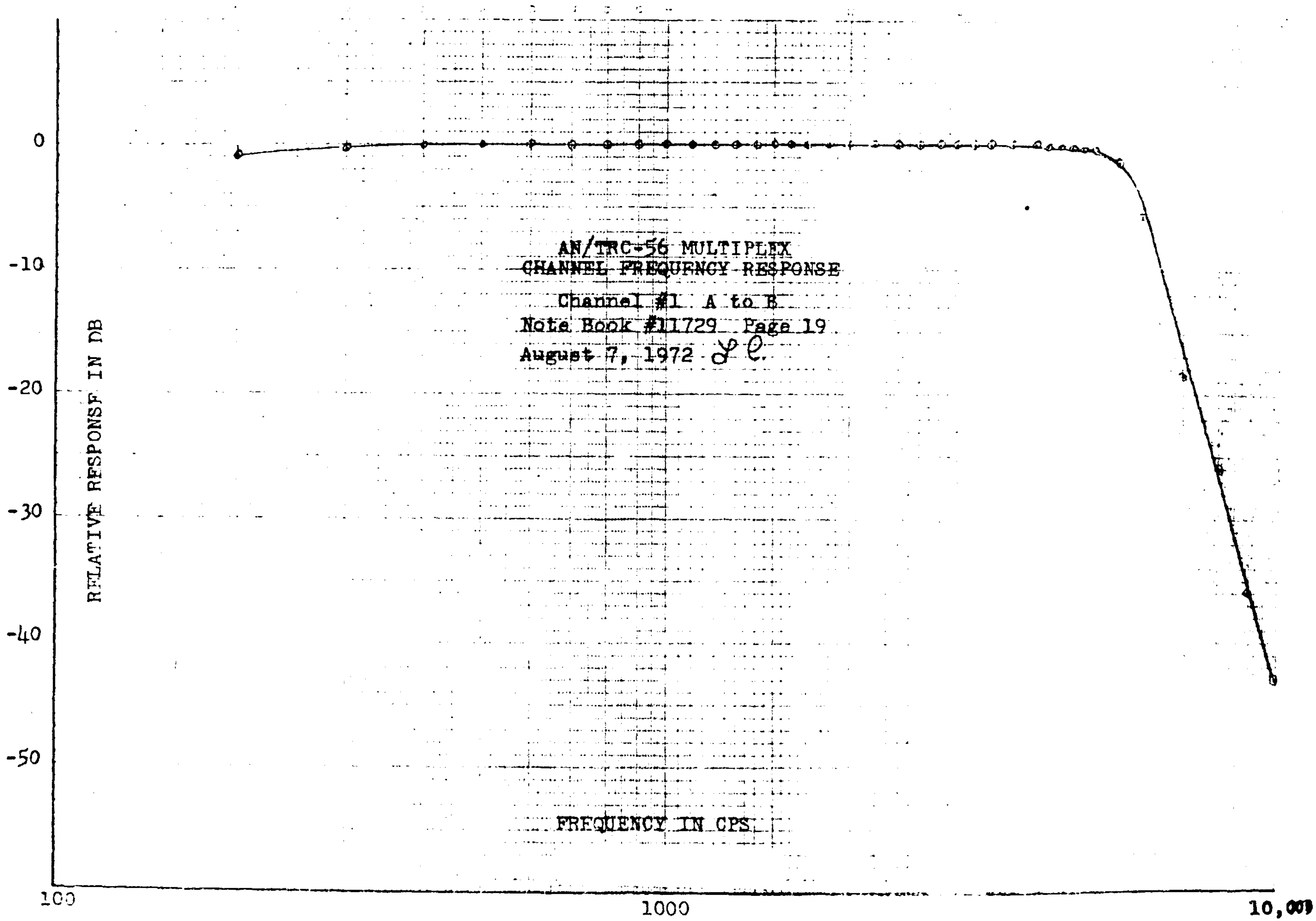


Figure 3 AN/TRC-56 Multiplex Channel Frequency Response, Channel 1, A to B

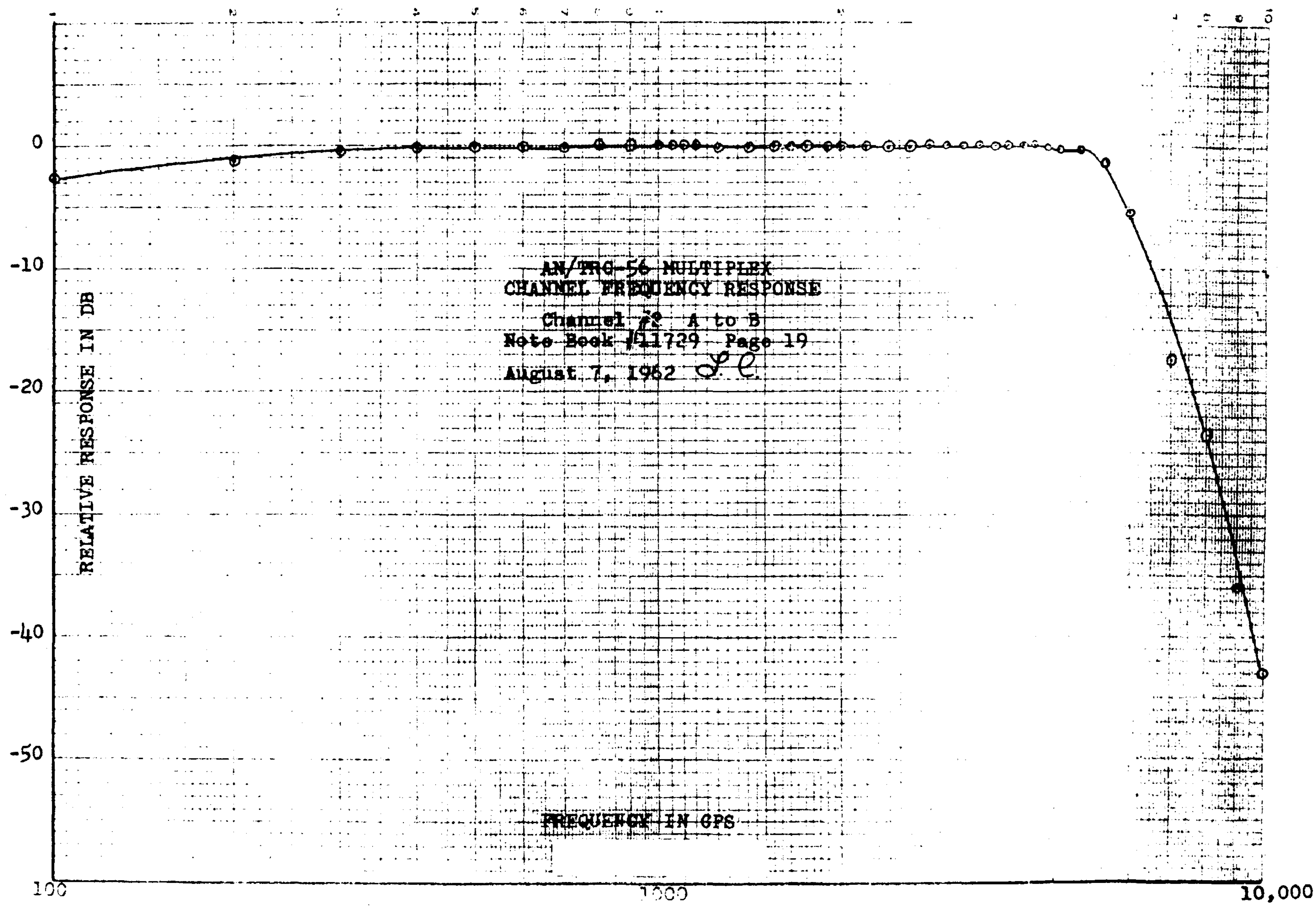


Figure 4 AN/TRC-56 Multiplex Channel Frequency Response, Channel 2, A to B

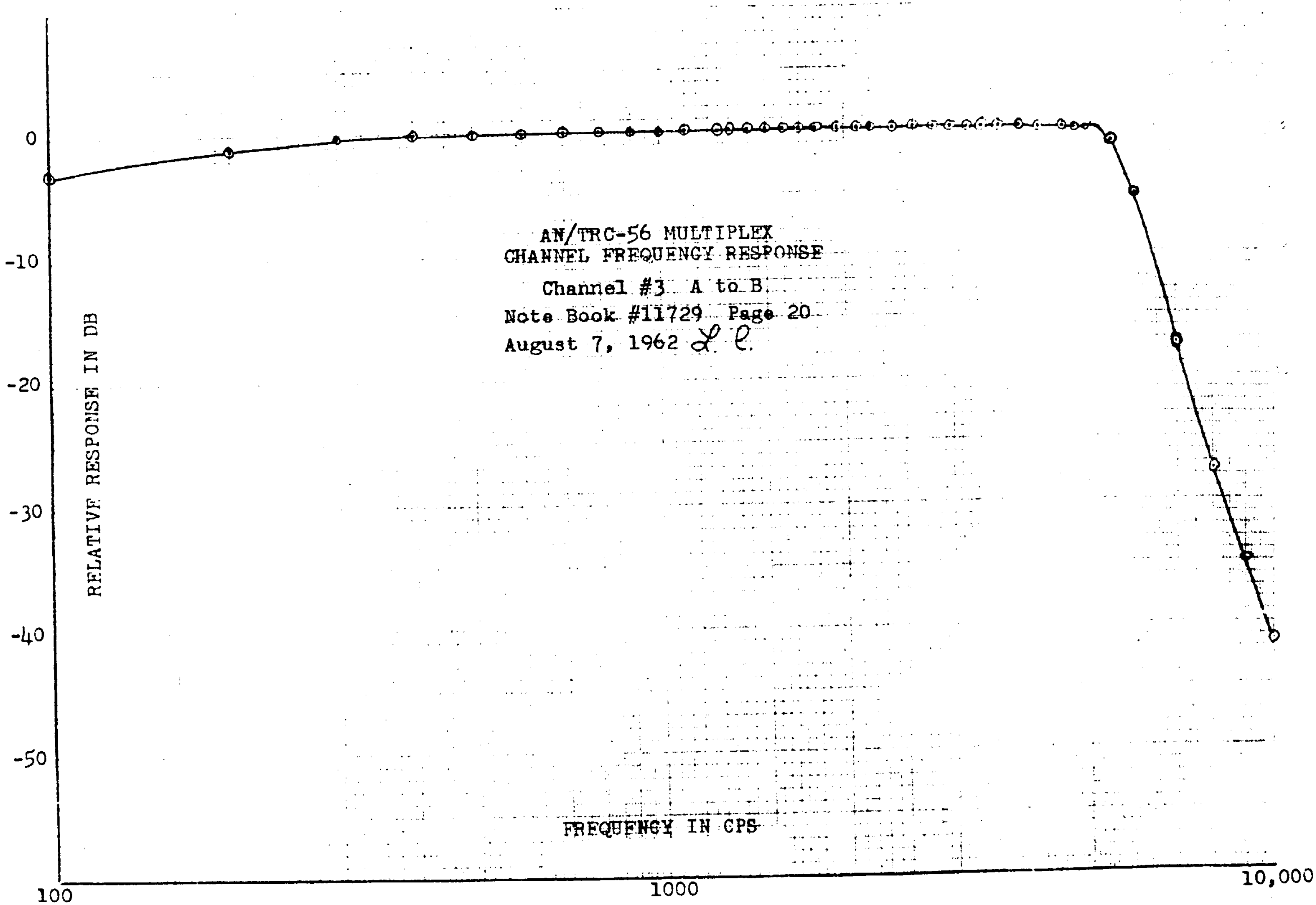


Figure 5 AN/TRC-56 Multiplex Channel Frequency Response, Channel 3, A to B

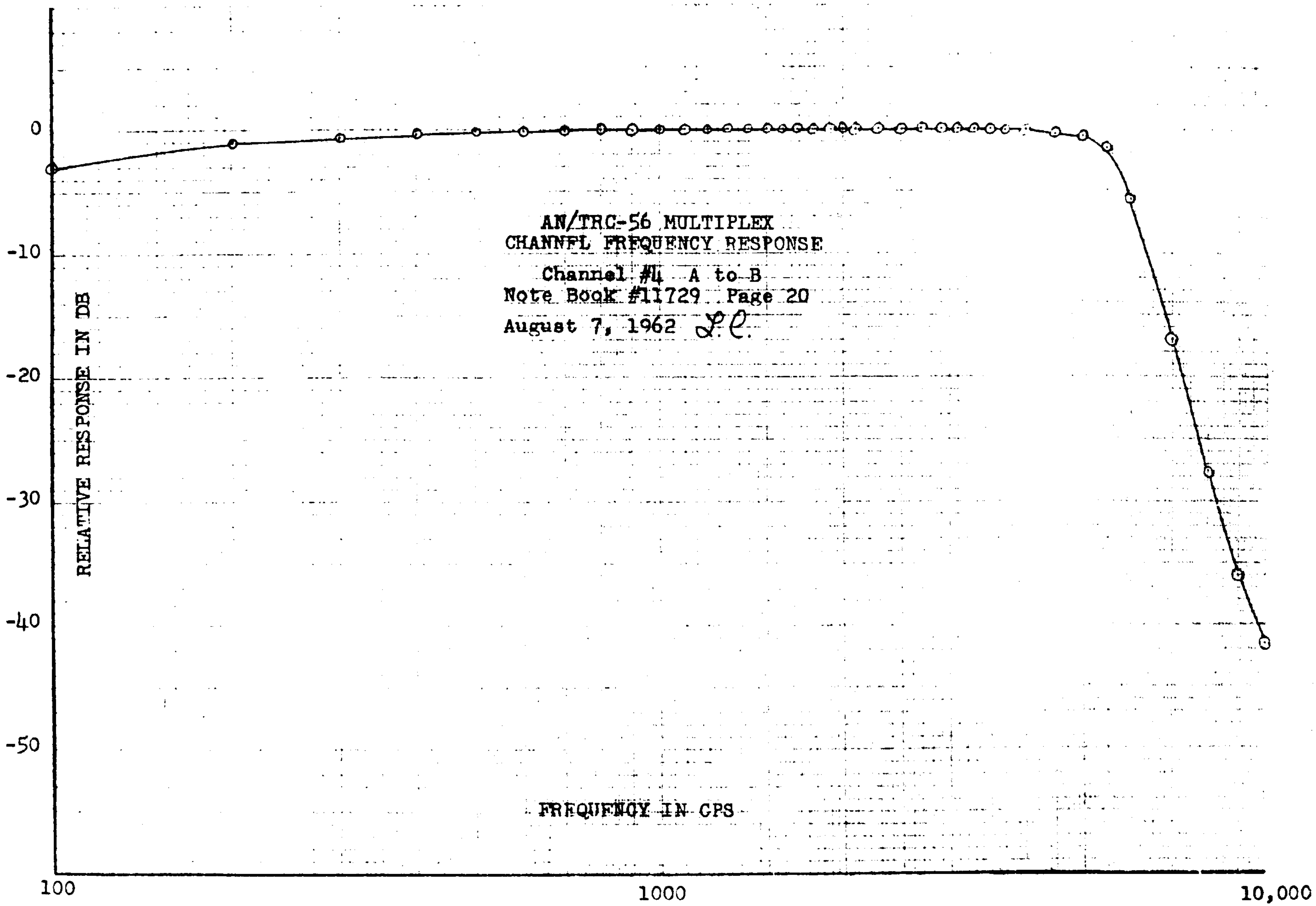


Figure 6 AN/TRC-56 Multiplex Channel Frequency Response, Channel 4, A to B

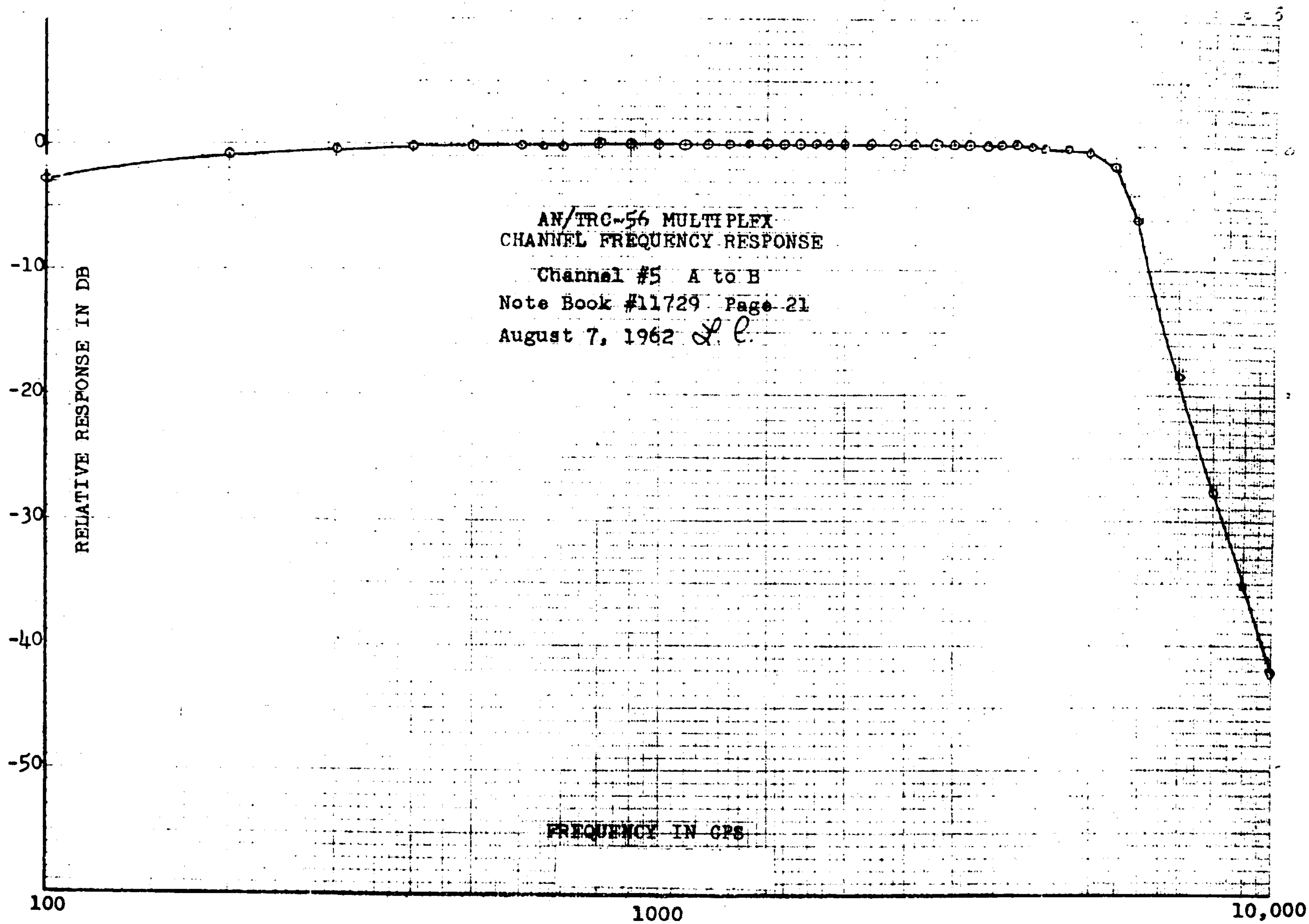


Figure 7 AN/TRC-56 Multiplex Channel Frequency Response, Channel 5, A to B

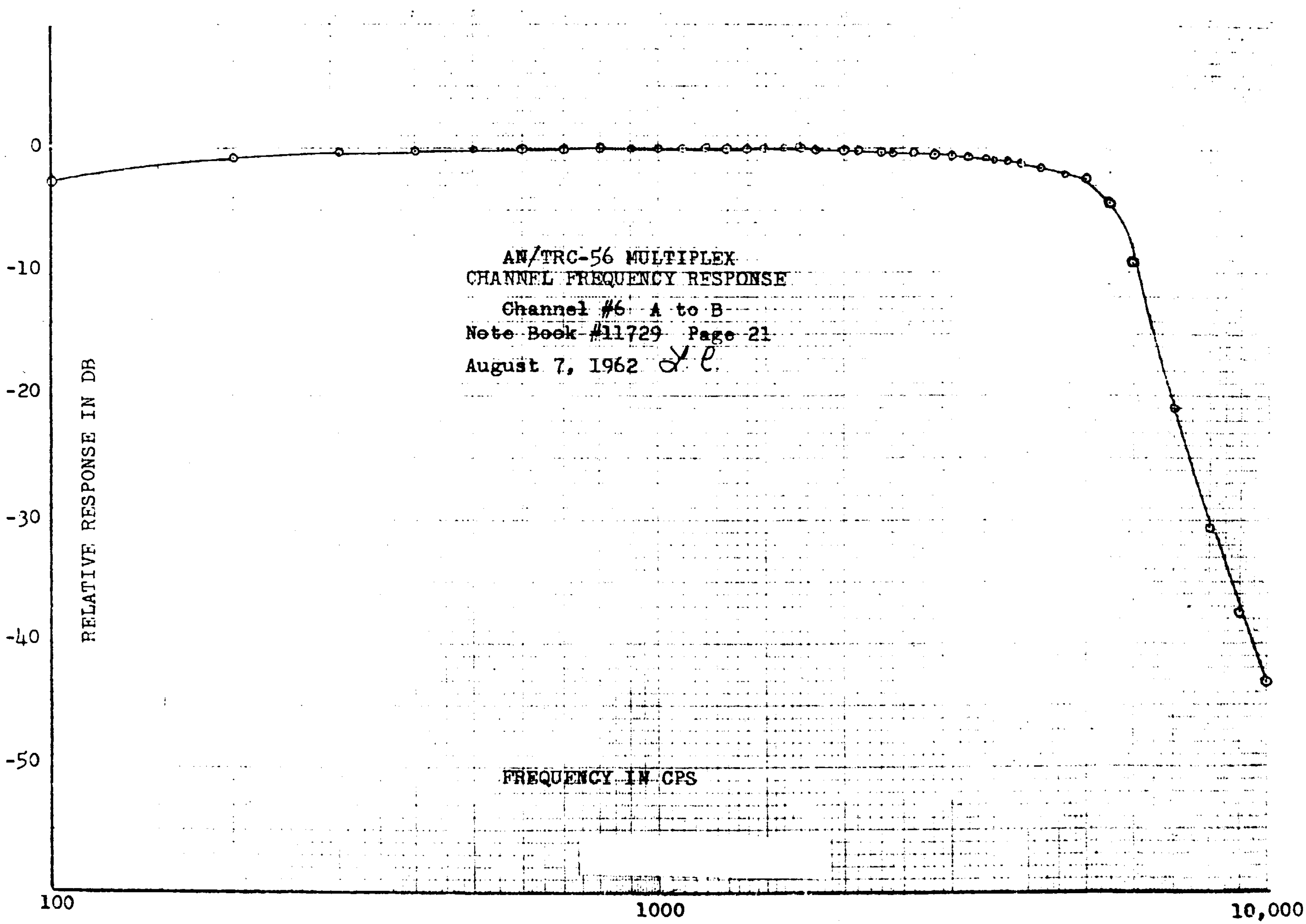


Figure 8 AN/TRC-56 Multiplex Channel Frequency Response, Channel 6, A to B

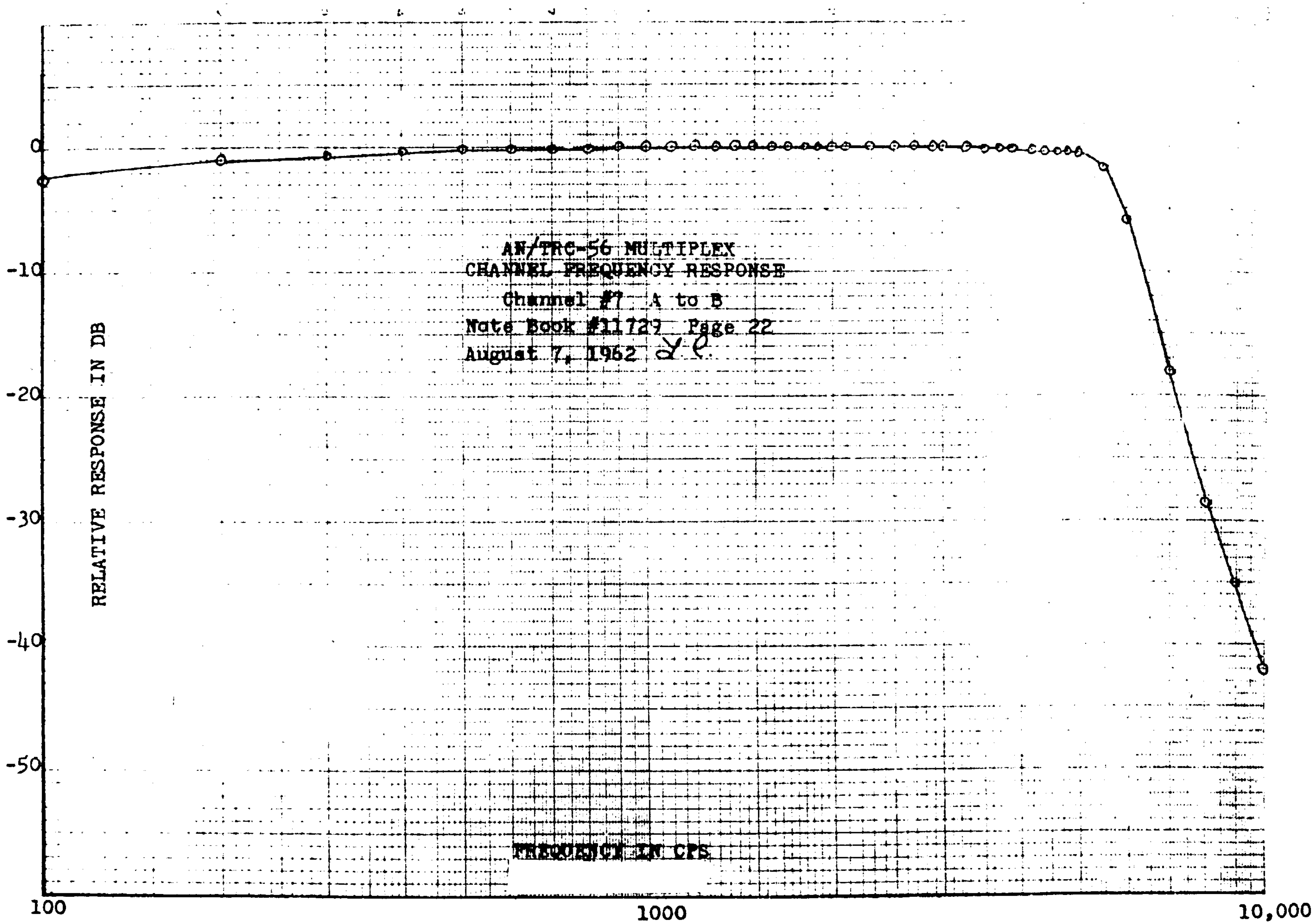


Figure 9 AN/TRC-56 Multiplex Channel Frequency Response, Channel 7, A to B

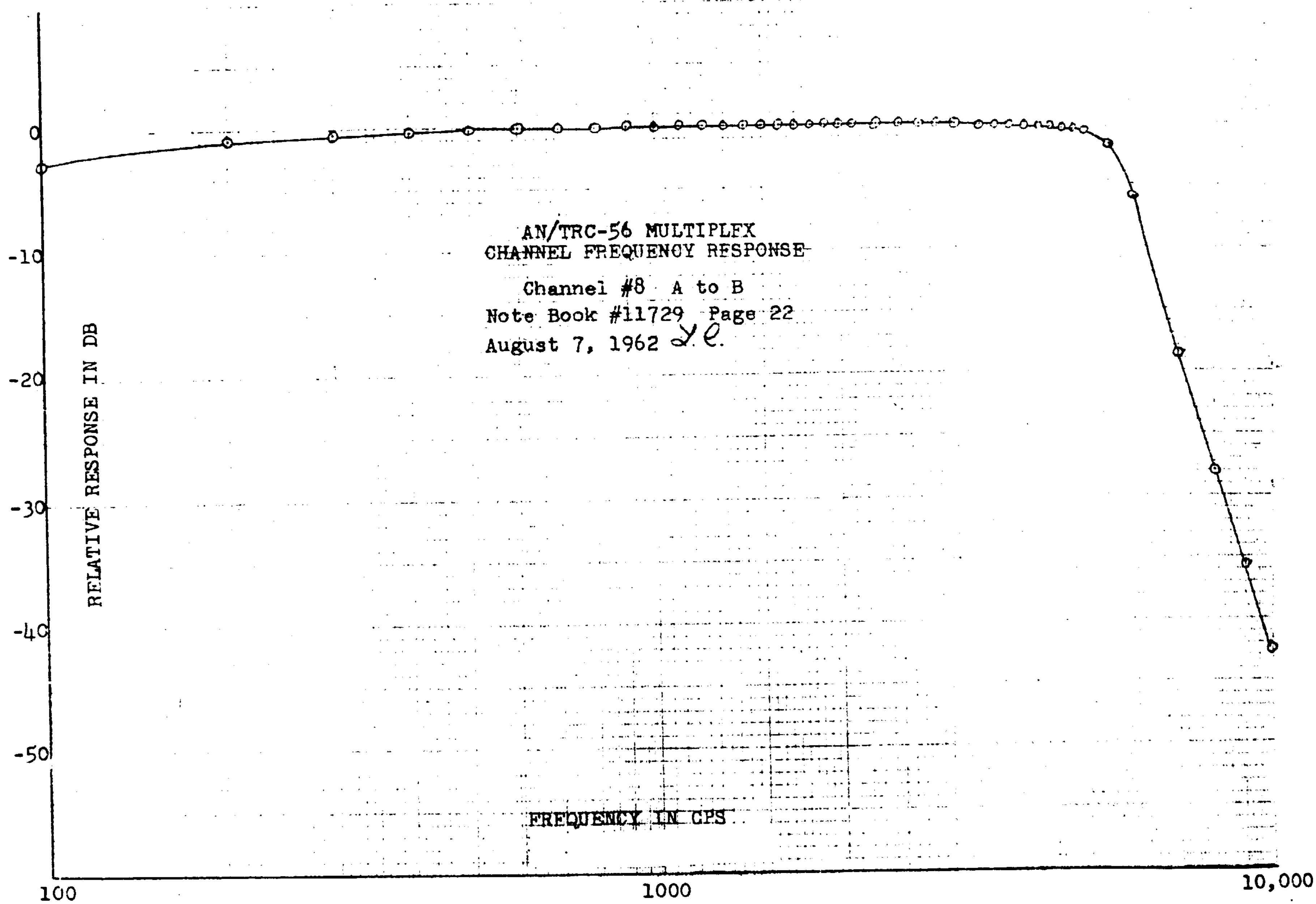


Figure 10 AN/TRC-56 Multiplex Channel Frequency Response, Channel 8, A to B

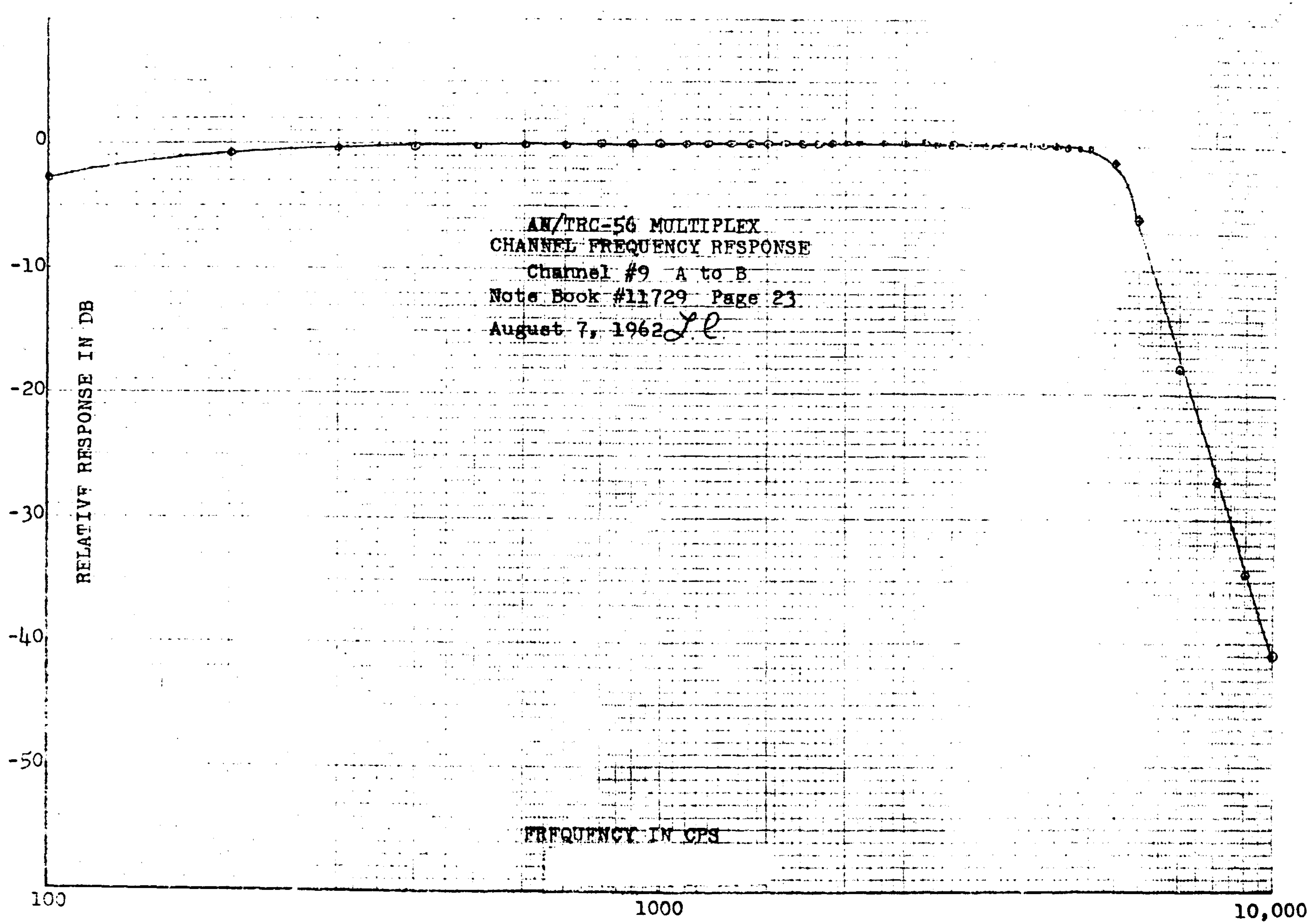


Figure 11 AN/TRC-56 Multiplex Channel Frequency Response, Channel 9, A to B

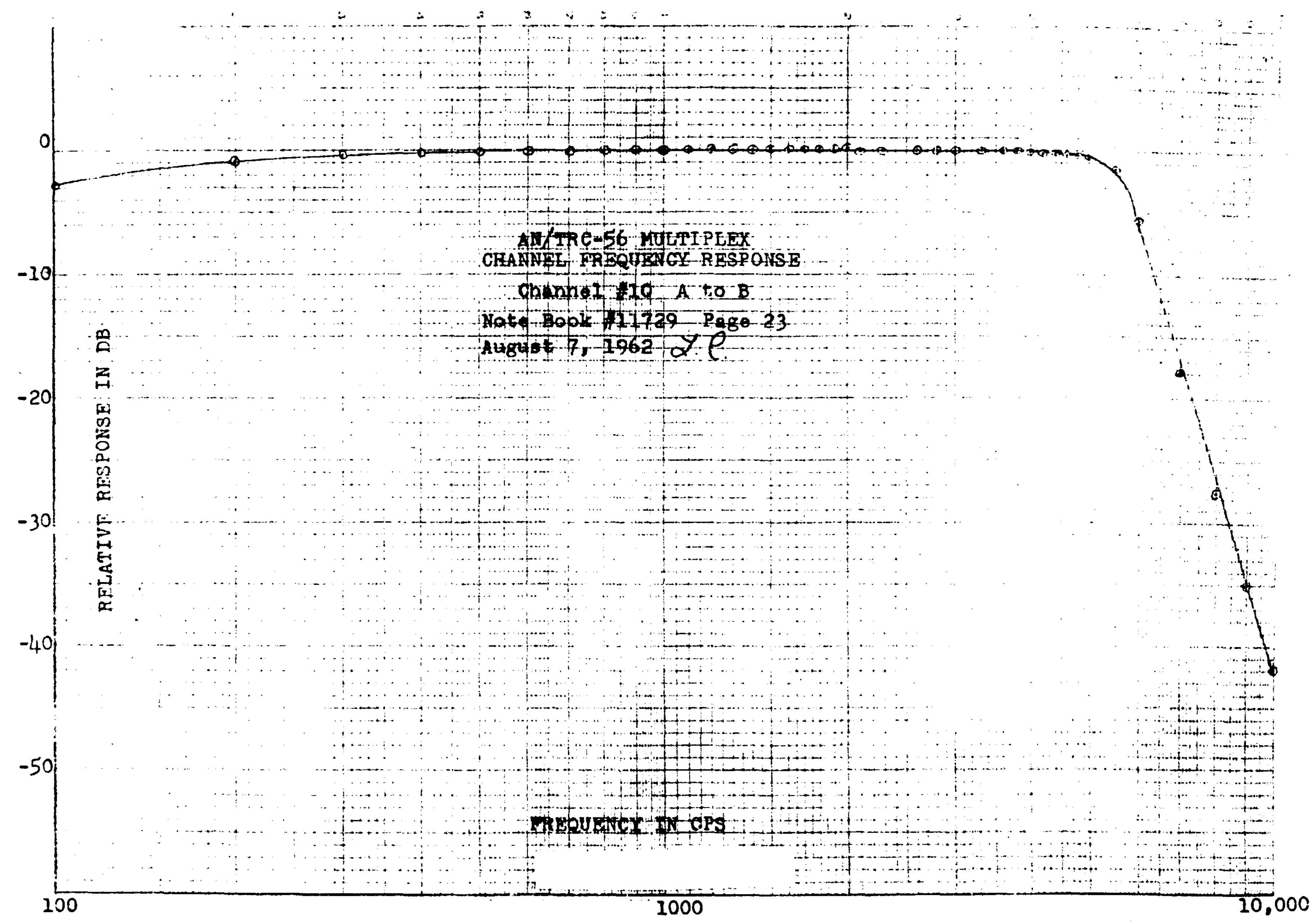


Figure 12 AN/TRC-56 Multiplex Channel Frequency Response, Channel 10, A to B

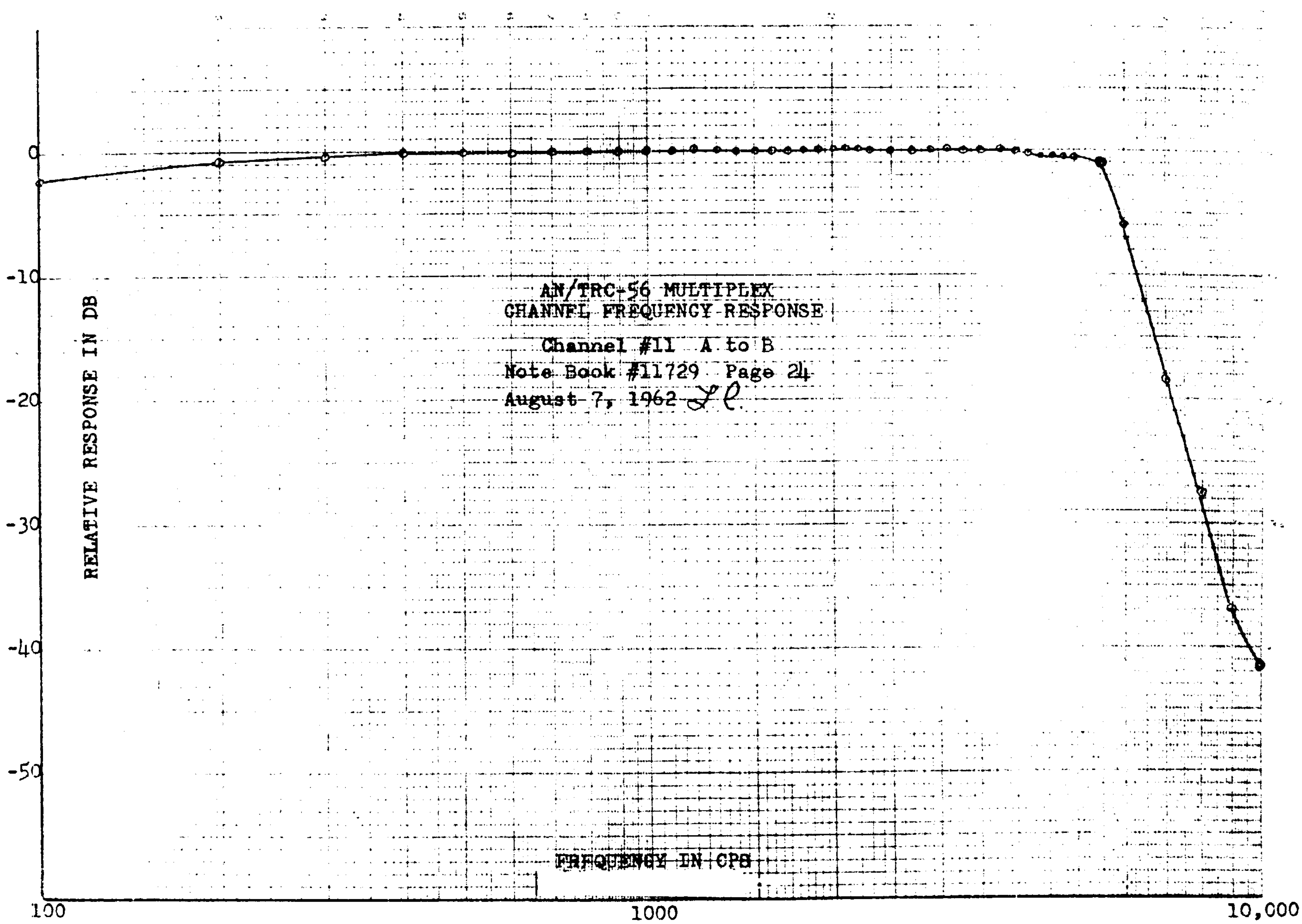


Figure 13 AN/TRC-56 Multiplex Channel Frequency Response, Channel 11, A to B

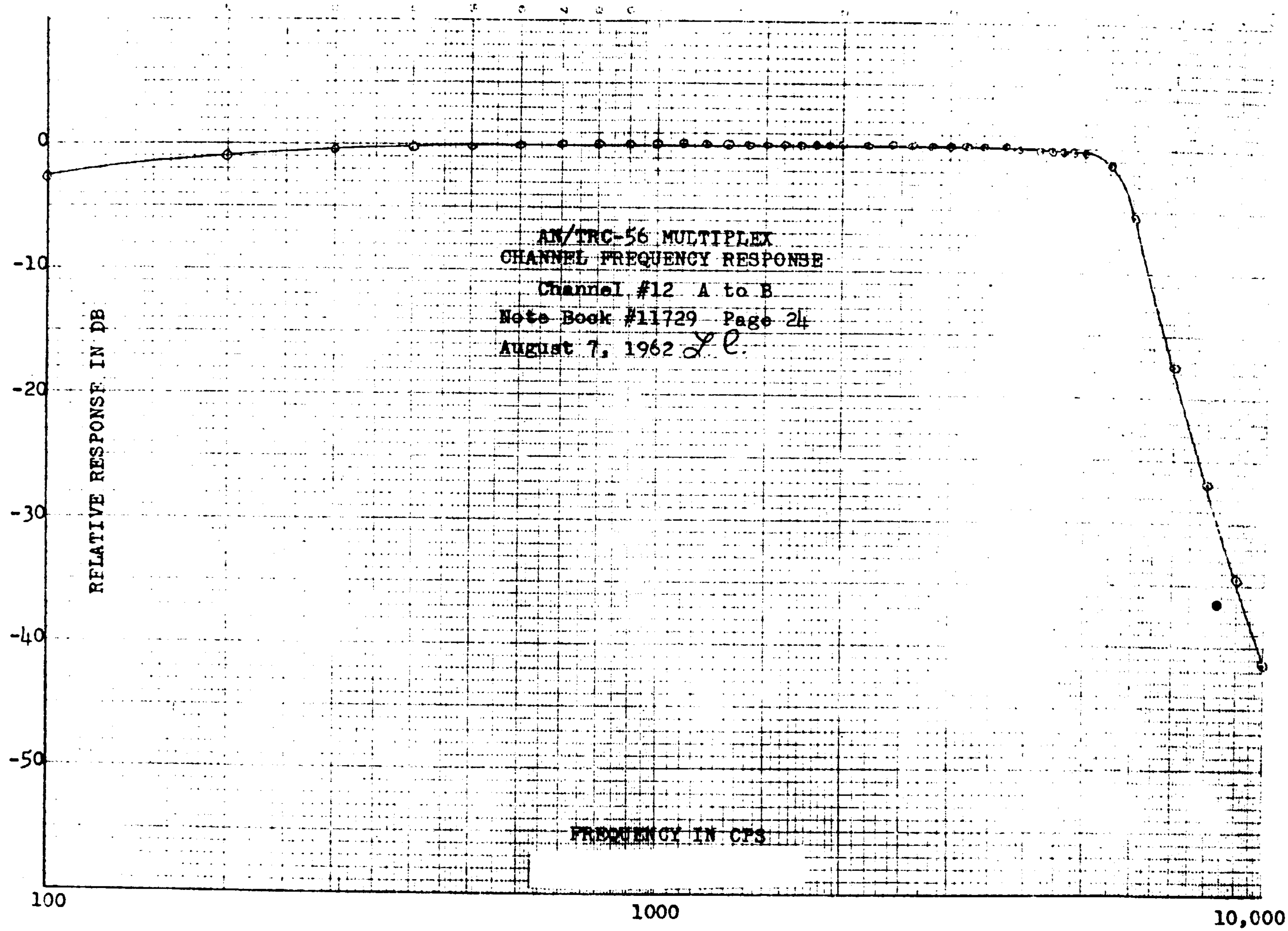


Figure 14 AN/TRC-56 Multiplex Channel Frequency Response, Channel 12, A to B

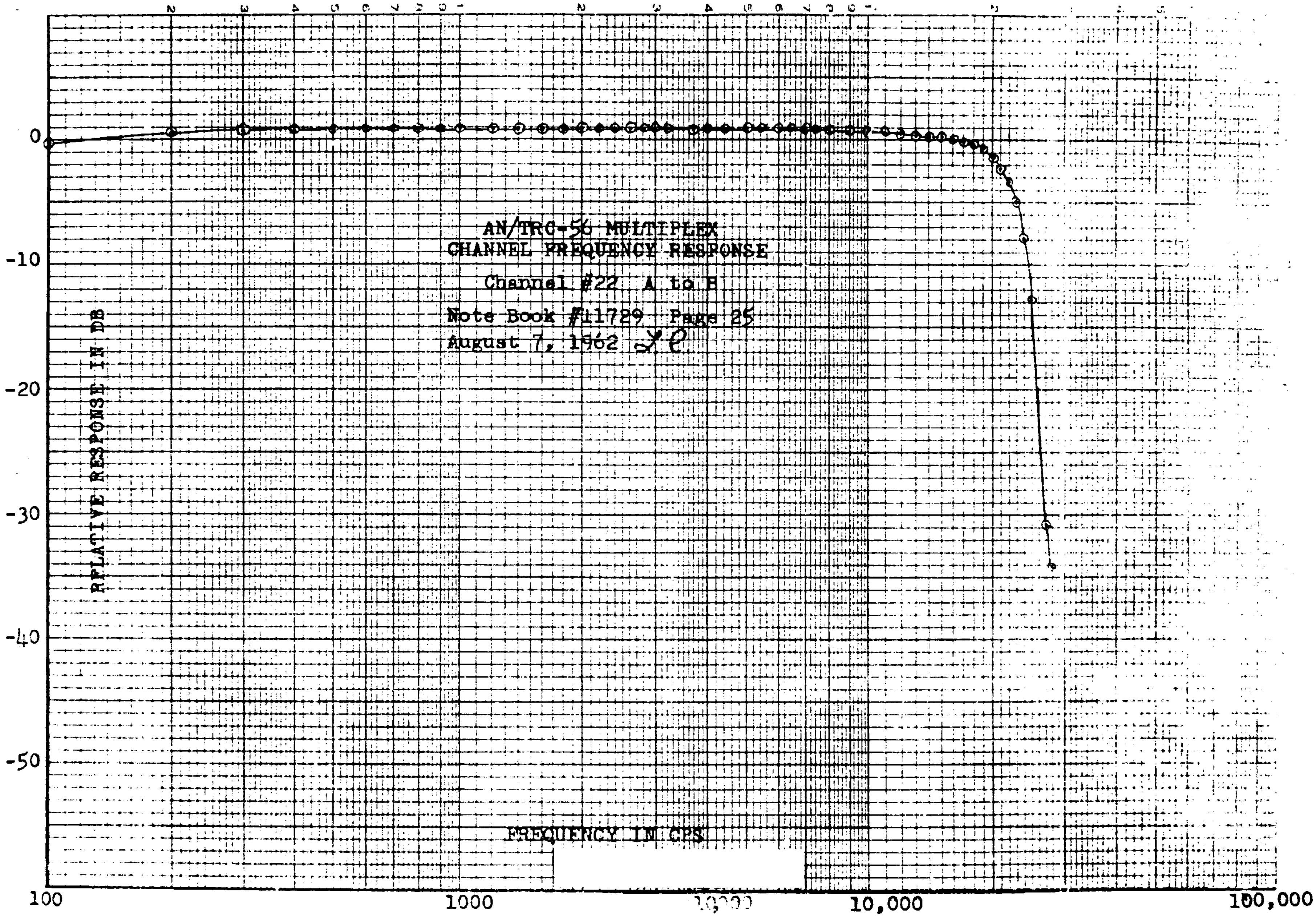


Figure 15 AN/TRC-56 Multiplex Channel Frequency Response, Channel 22, A to B

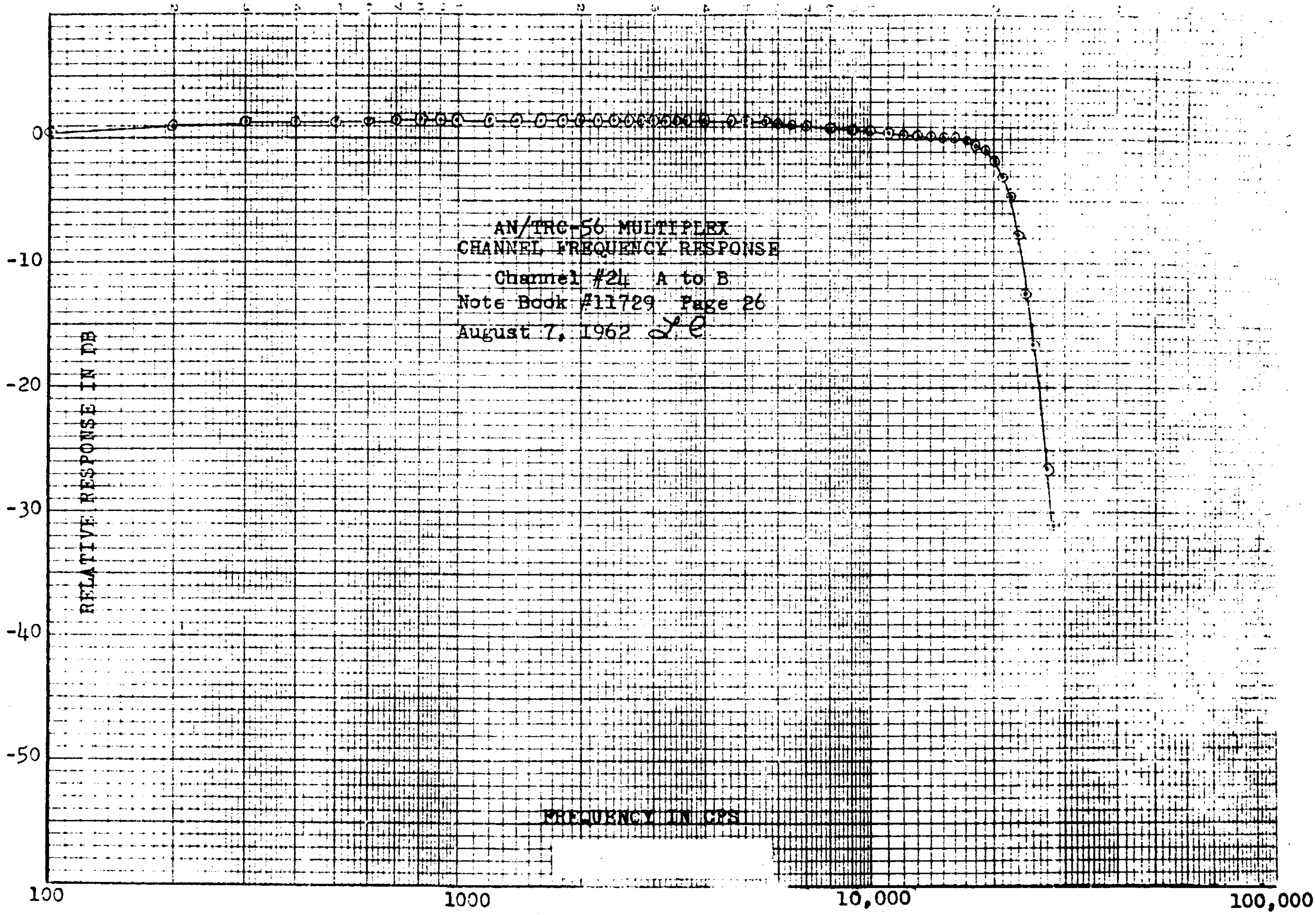


Figure 16 AN/TRC-56 Multiplex Channel Frequency Response, Channel 24, A to B

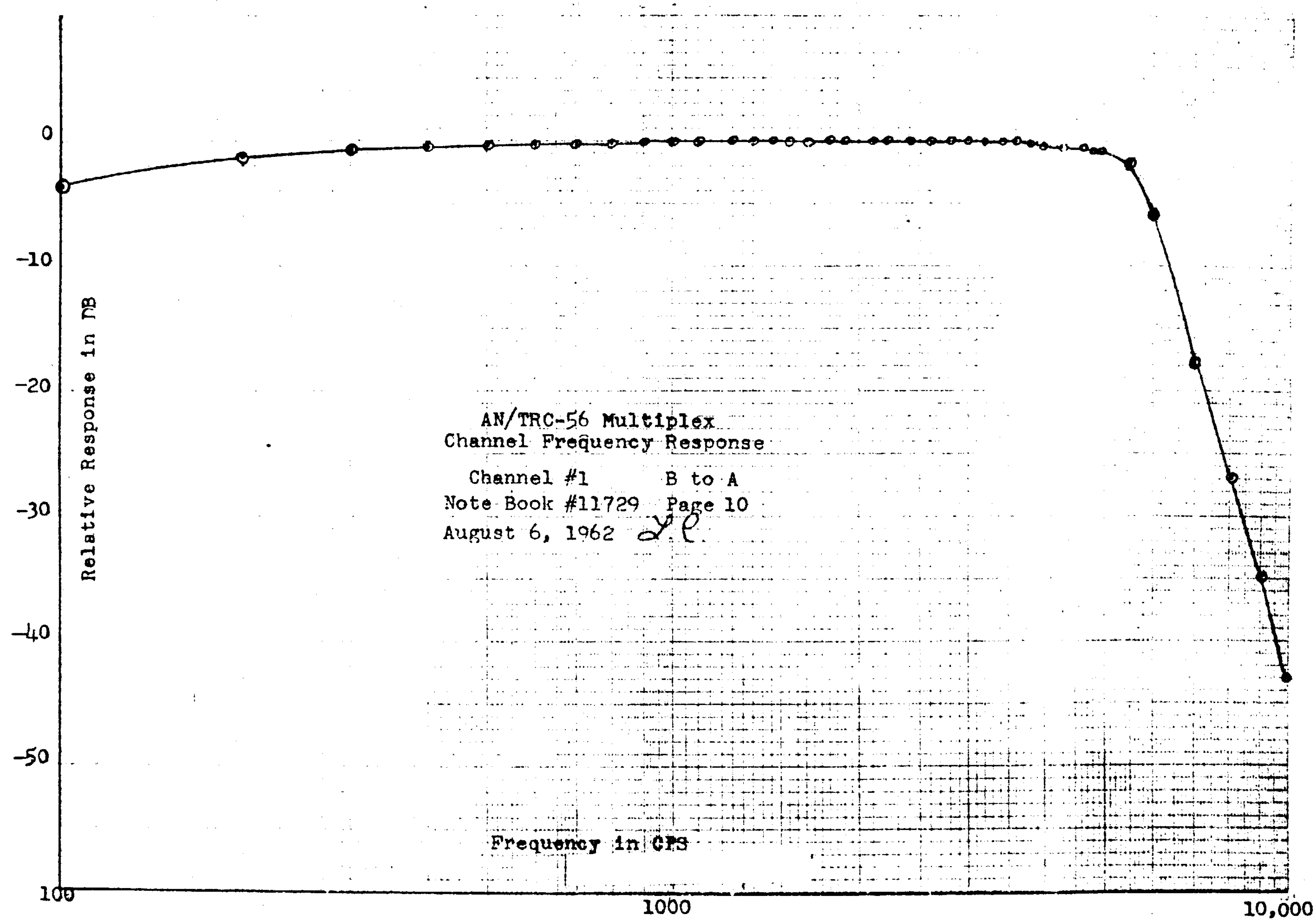


Figure 17 AN/TRC-56 Multiplex Channel Frequency Response, Channel 1, B to A

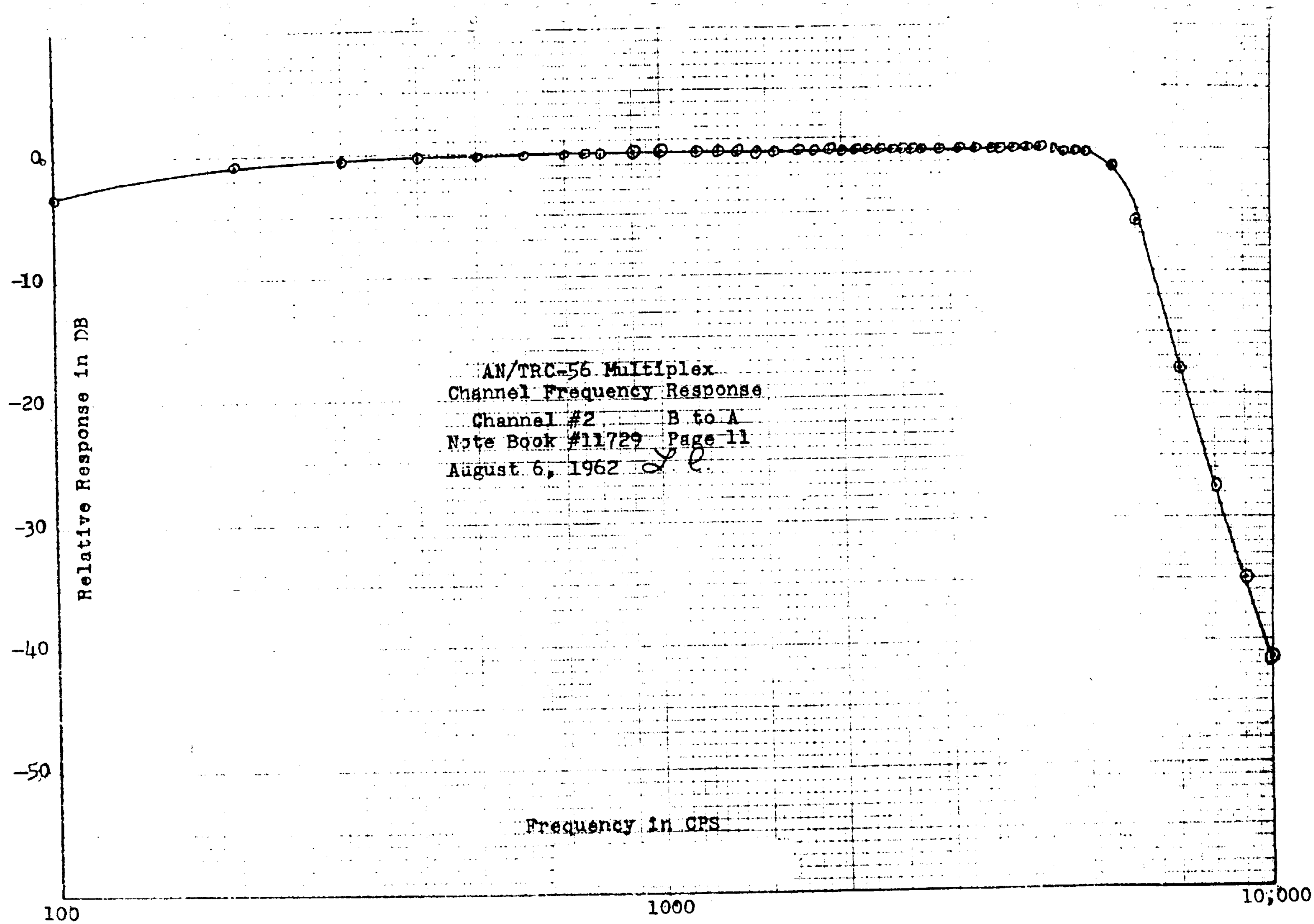


Figure 18 AN/TRC-56 Multiplex Channel Frequency Response, Channel 2, 3 to A

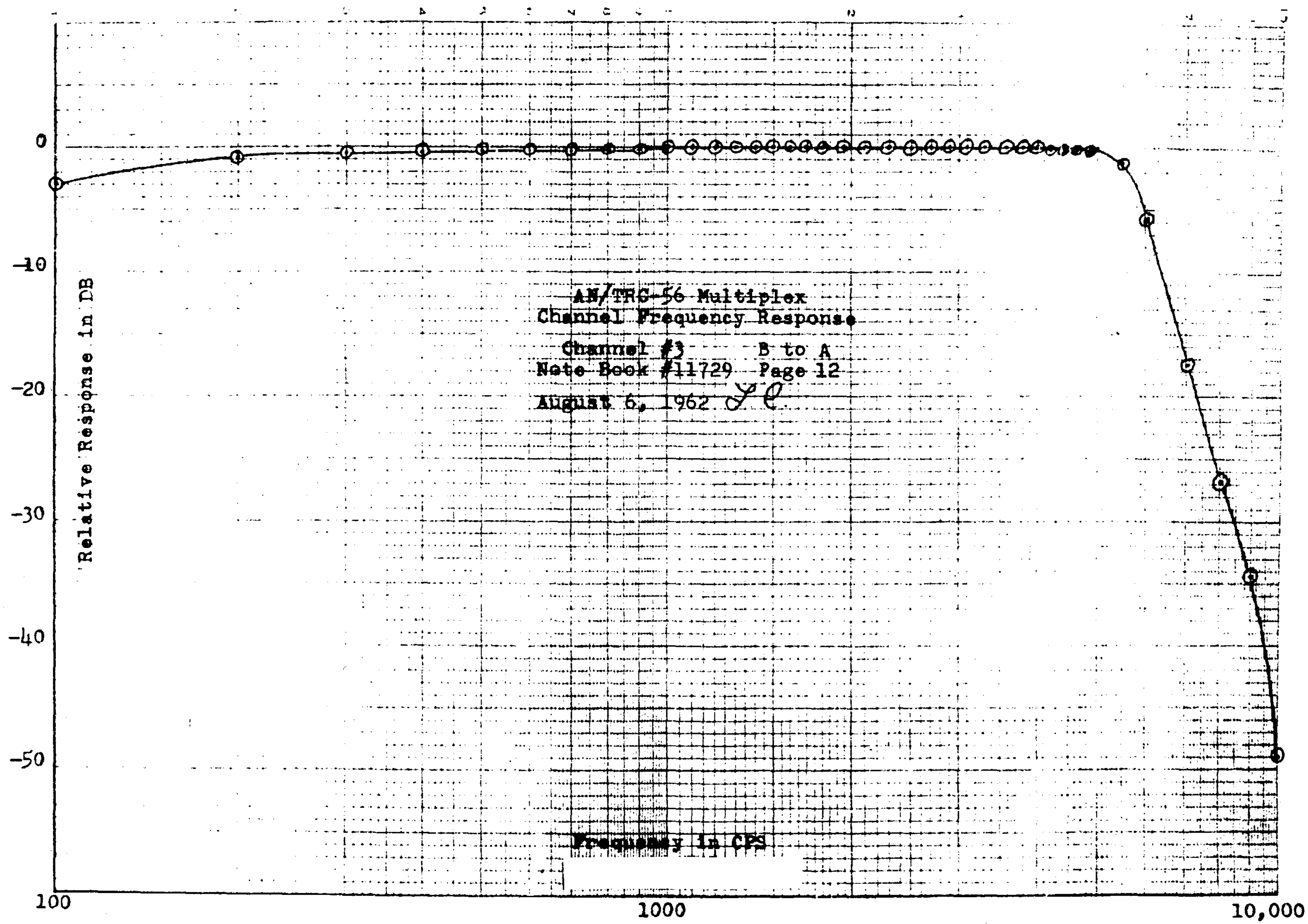


Figure 19 AN/TRC-56 Multiplex Channel Frequency Response, Channel 3, B to A

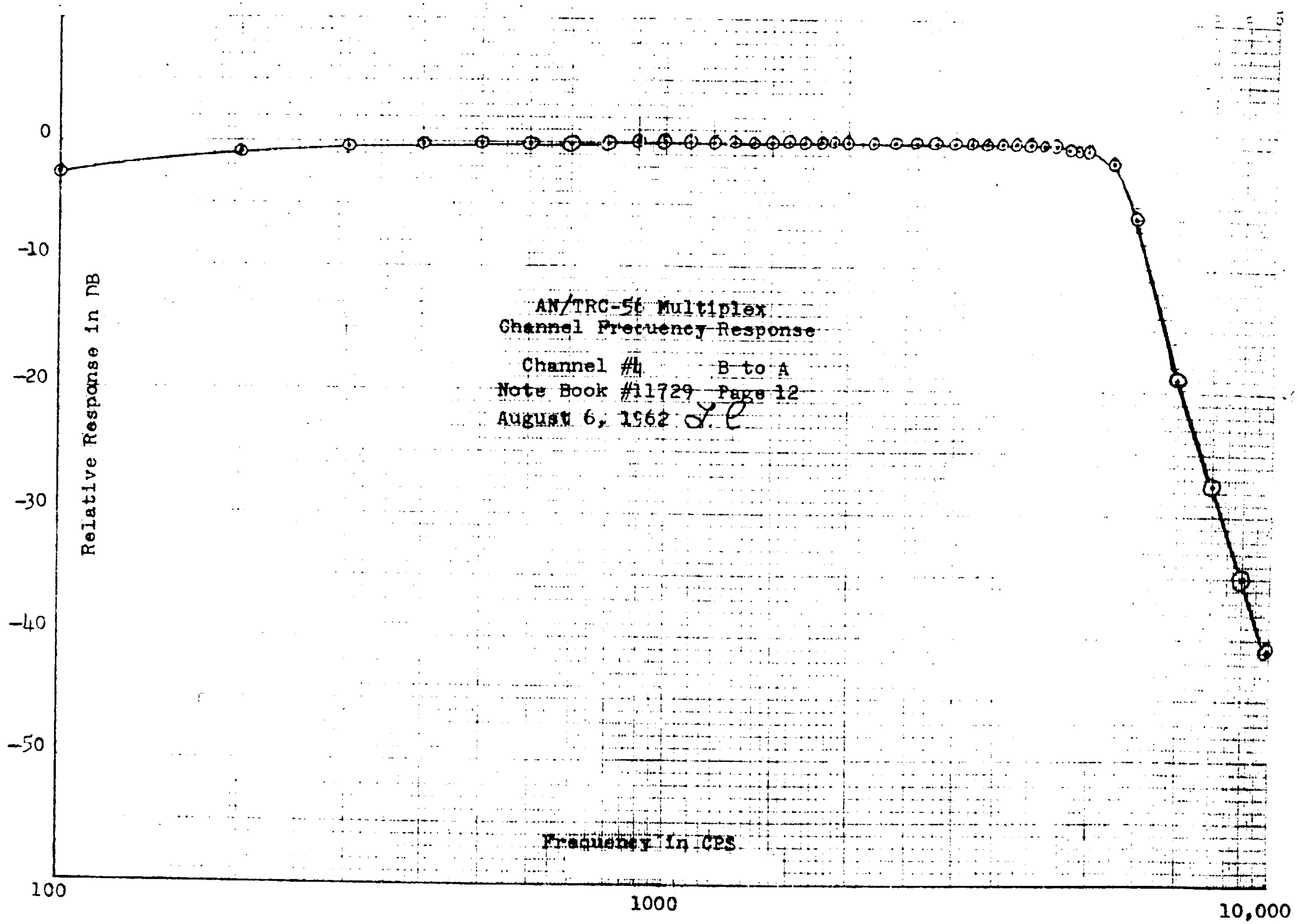


Figure 20 AN/TRC 56 Multiplex Channel Frequency Response, Channel 4, B to A

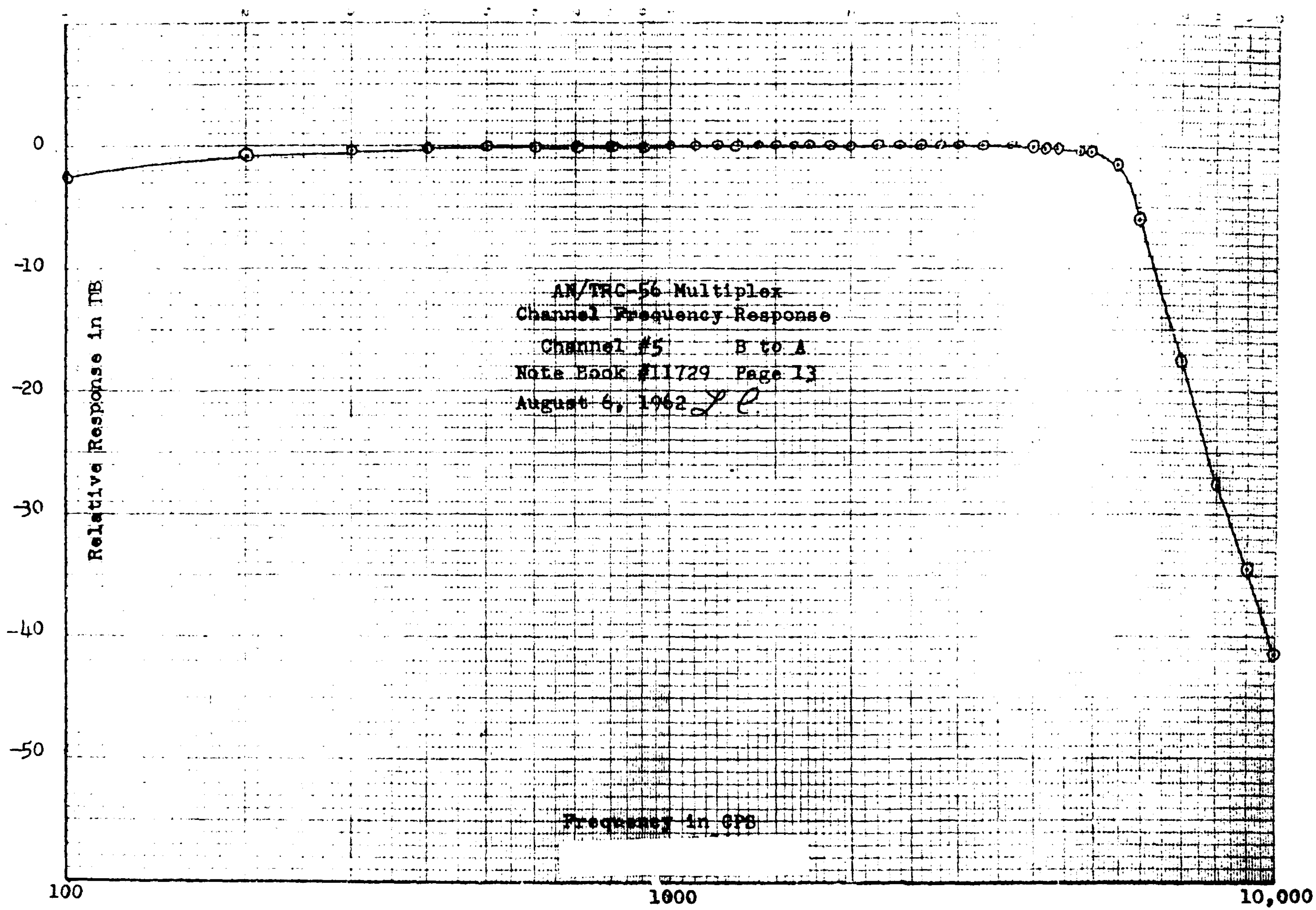


Figure 21 AN/TRC-56 Multiplex Channel Frequency Response, Channel 5, B to A

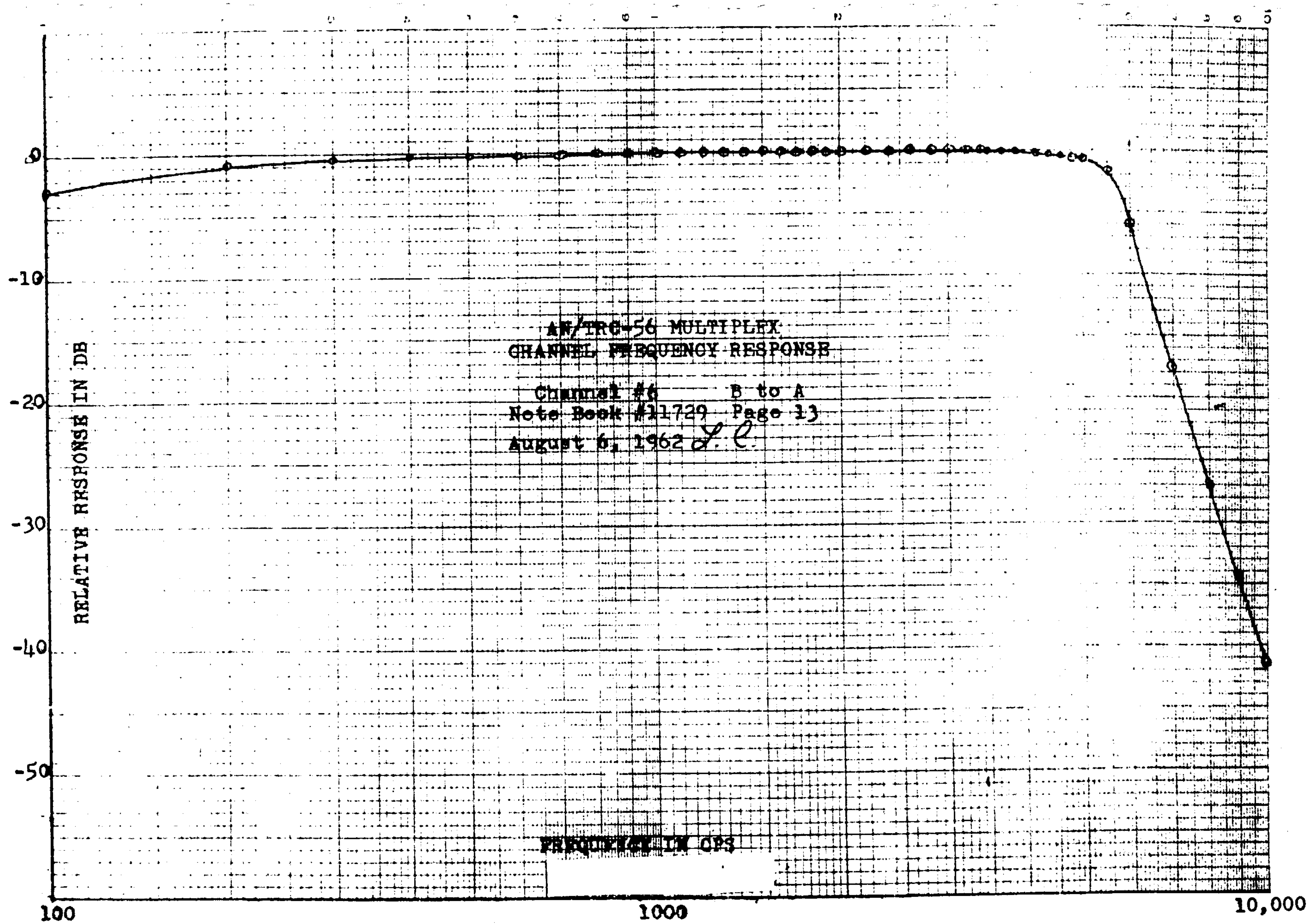


Figure 22 AN/TRC-56 Multiplex Channel Frequency Response, Channel 6, B to A

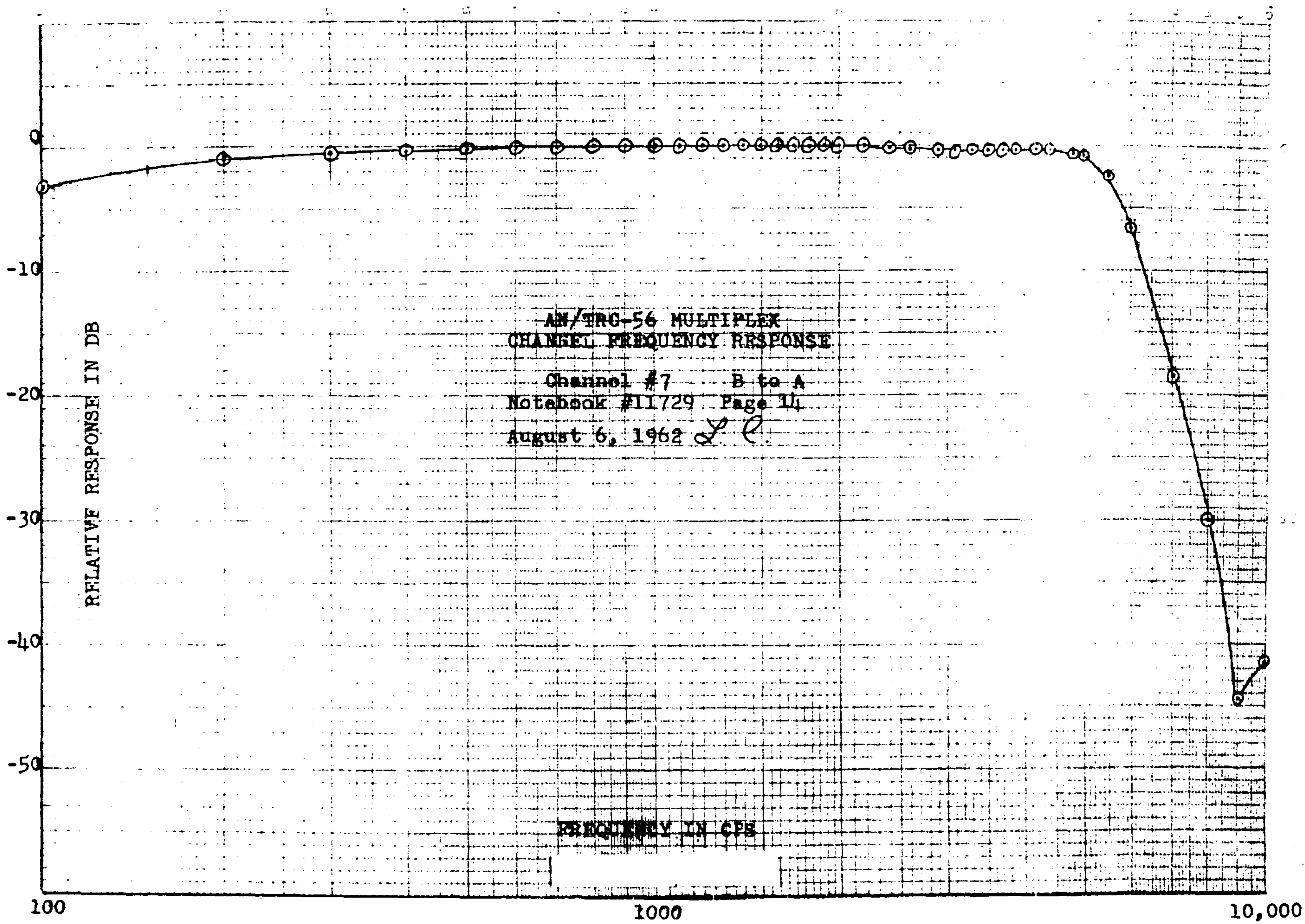


Figure 23 AN/TRC-56 Multiplex Channel Frequency Response, Channel 7, B to A

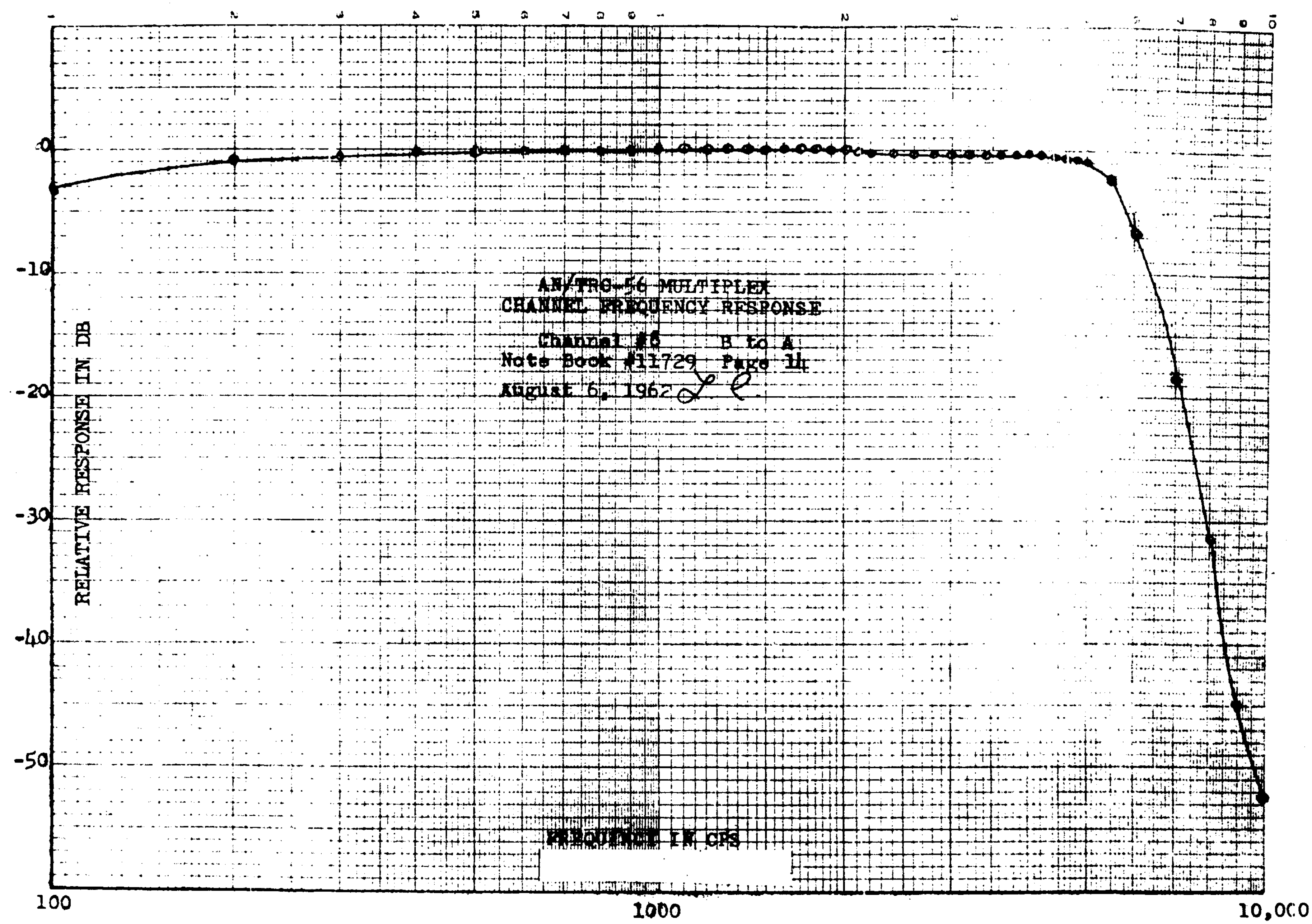


Figure 24 AN/TRC-56 Multiplex Channel Frequency Response, Channel 8, B to A

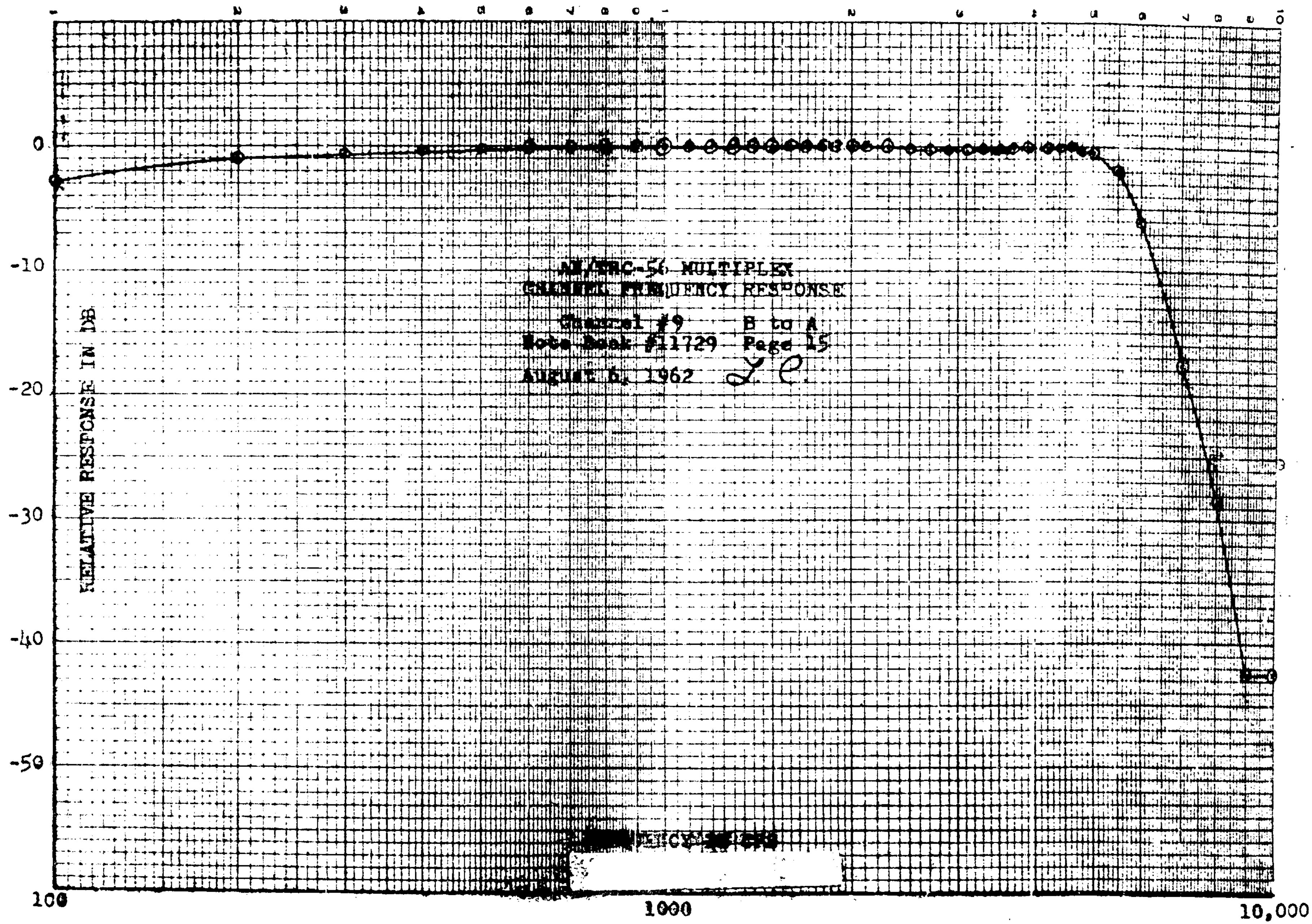


Figure 25 AN/TRC-56 Multiplex Channel Frequency Response, Channel 9 B to A

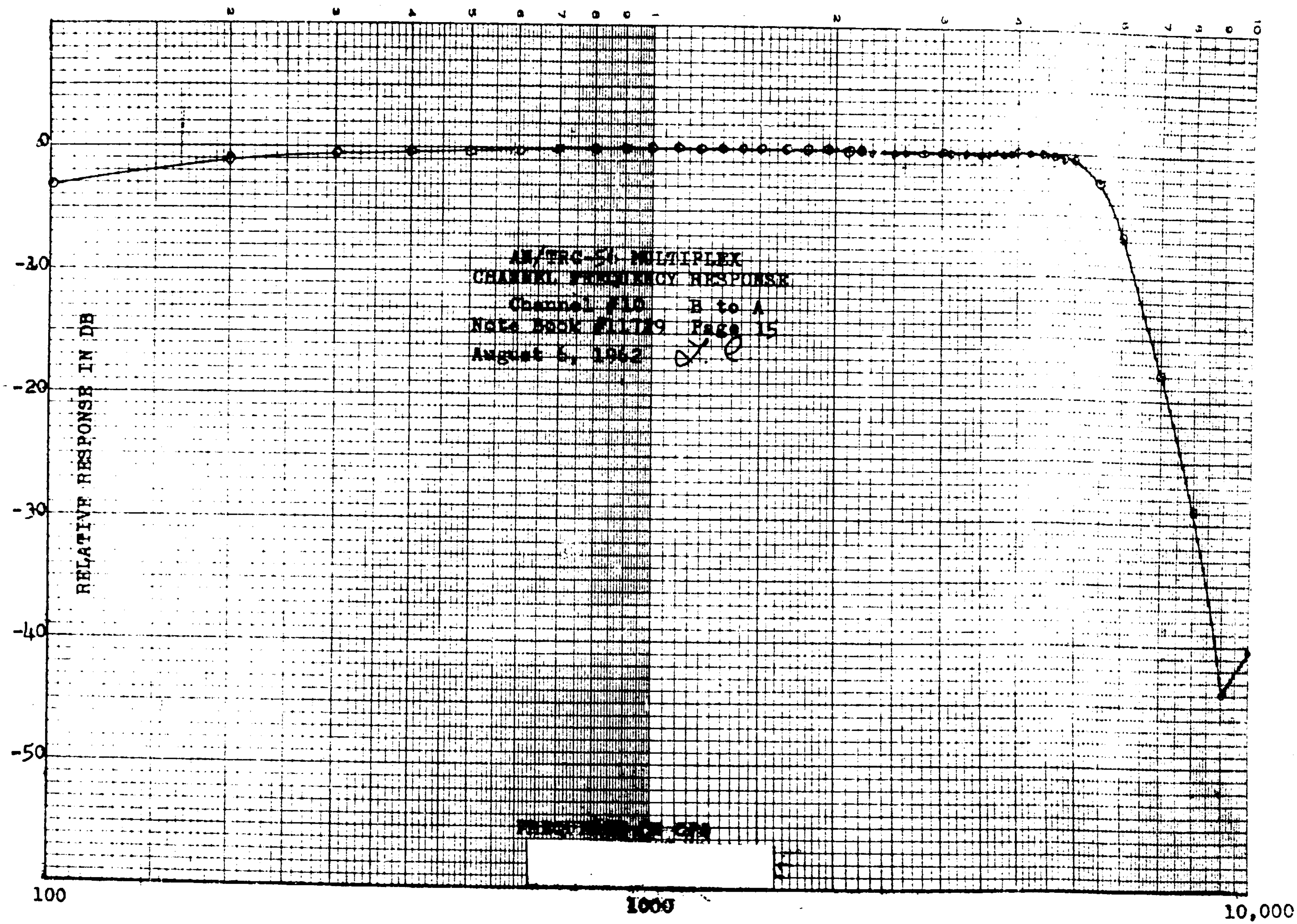


Figure 26 AN/TRC-56 Multiplex Channel Frequency Response, Channel 10, B to A

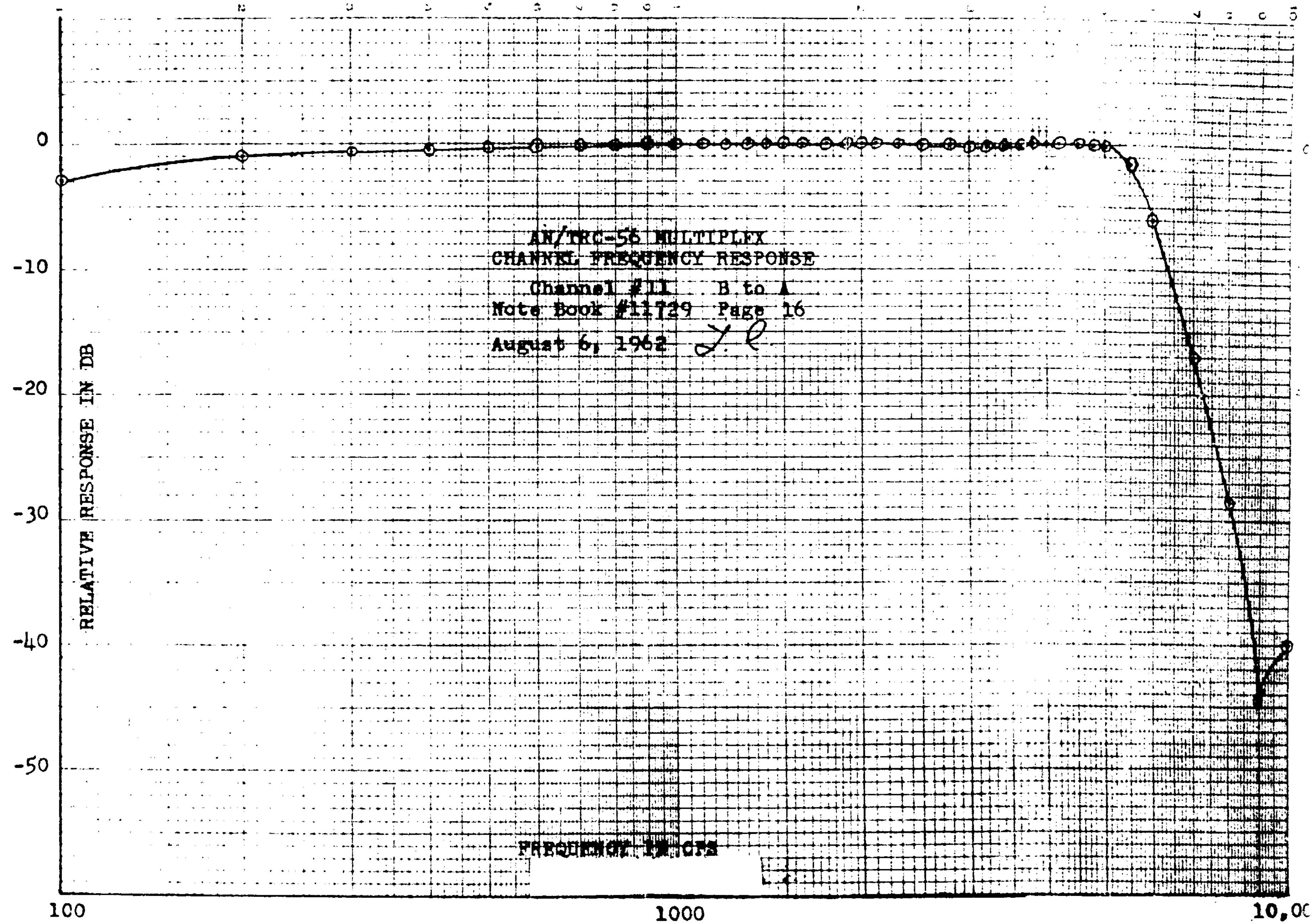


Figure 27 AN/TRC-56 Multiplex Channel Frequency Response, Channel 11, B to A

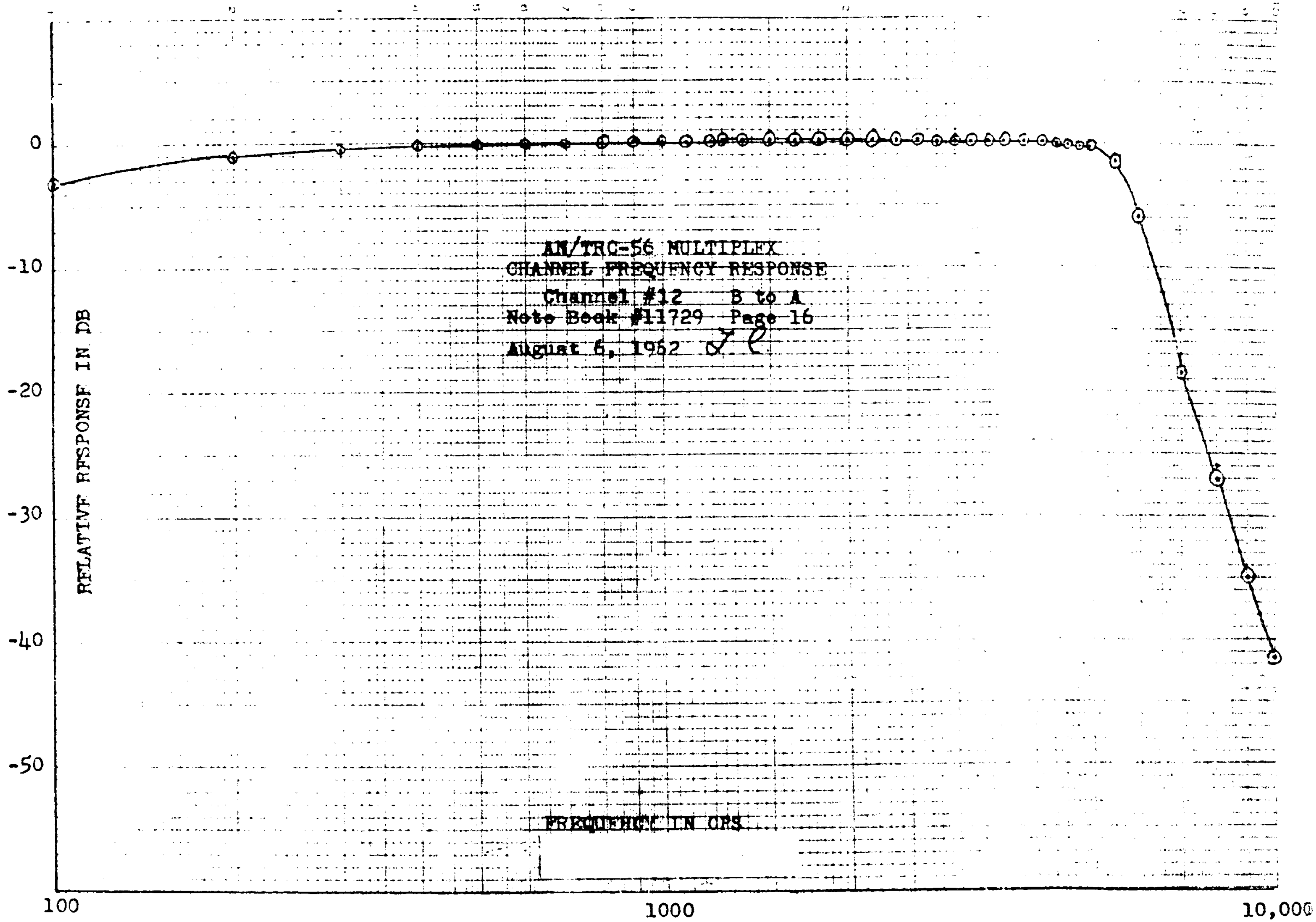


Figure 28 AN/TRC-56 Multiplex Channel Frequency Response, Channel 12, B to A

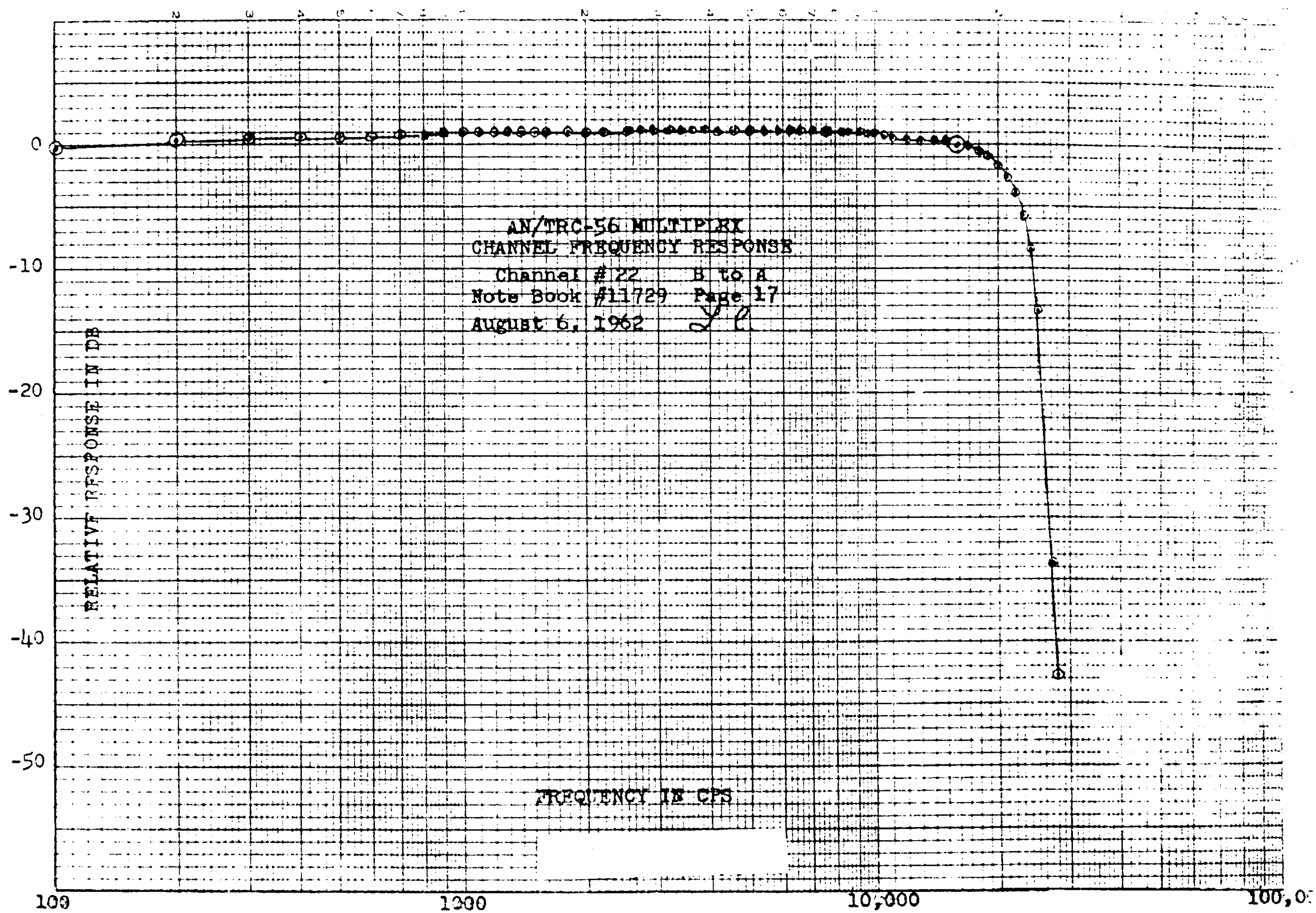


Figure 29 AN/TRC-56 Multiplex Channel Frequency Response, Channel 22, B to A

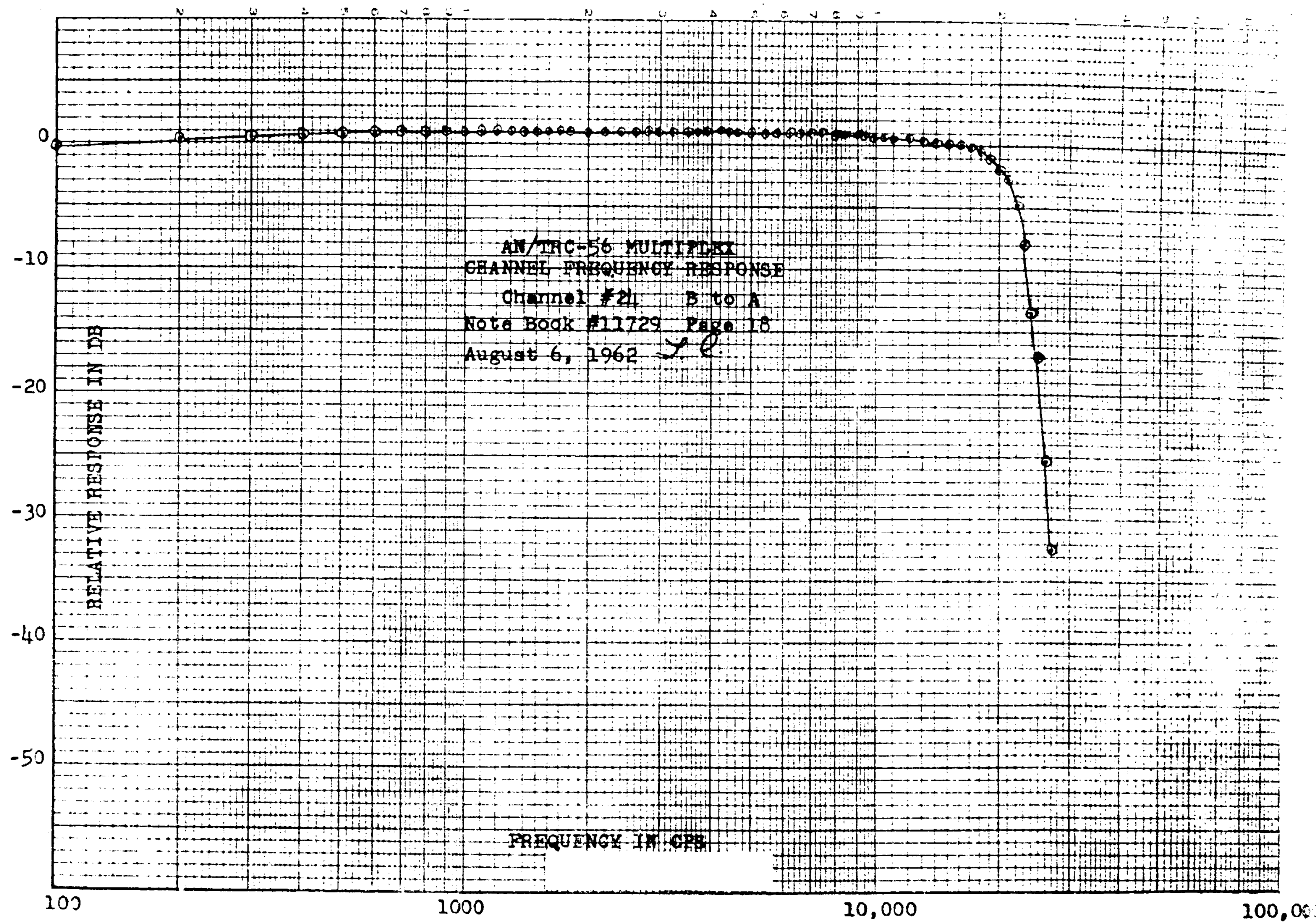


Figure 30 AN/TRC-56 Multiplex Channel Frequency Response, Channel 24, B to A

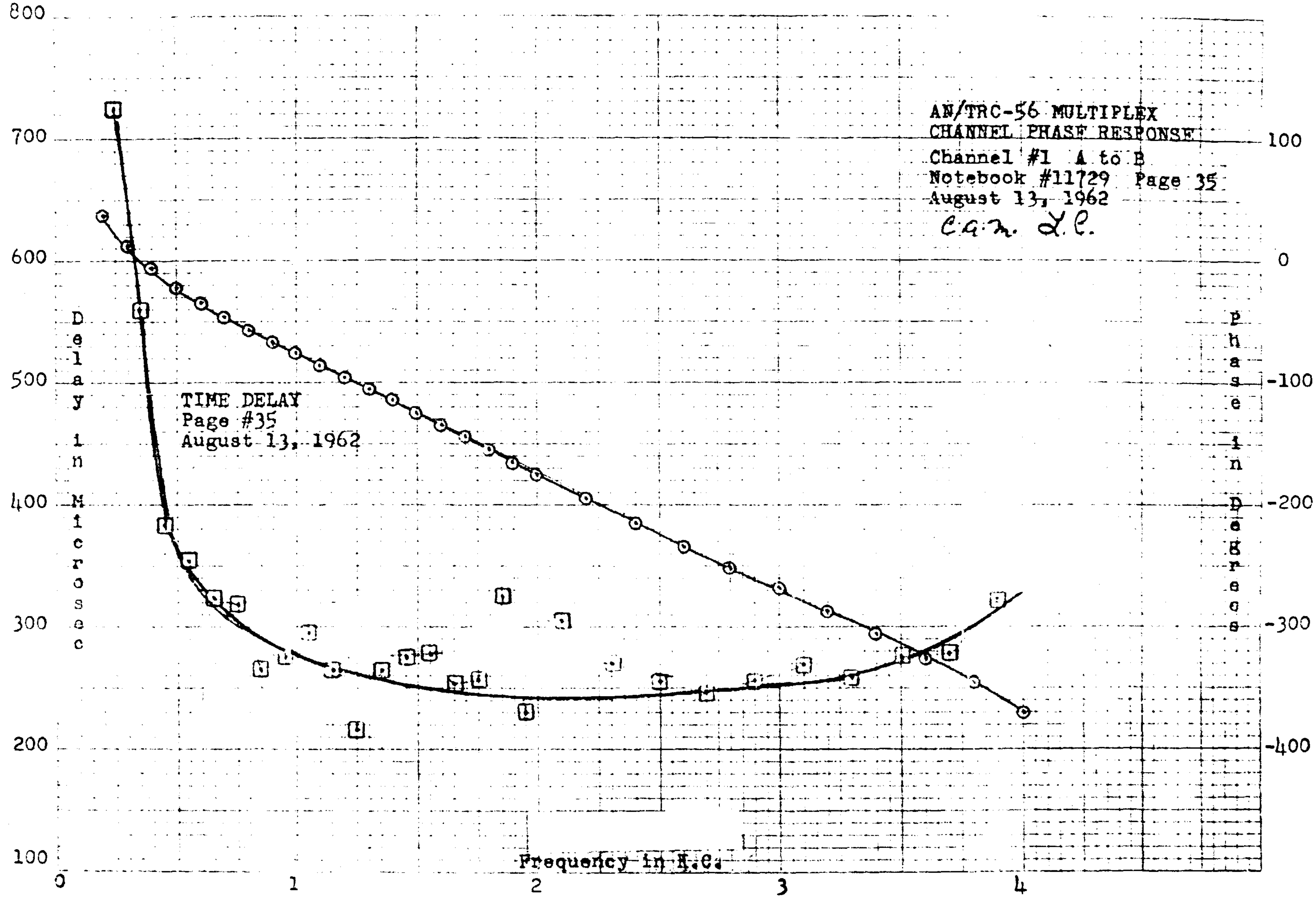


Figure 31 AN/TRC-56 Multiplex Channel Phase Response, Channel 1, A to B

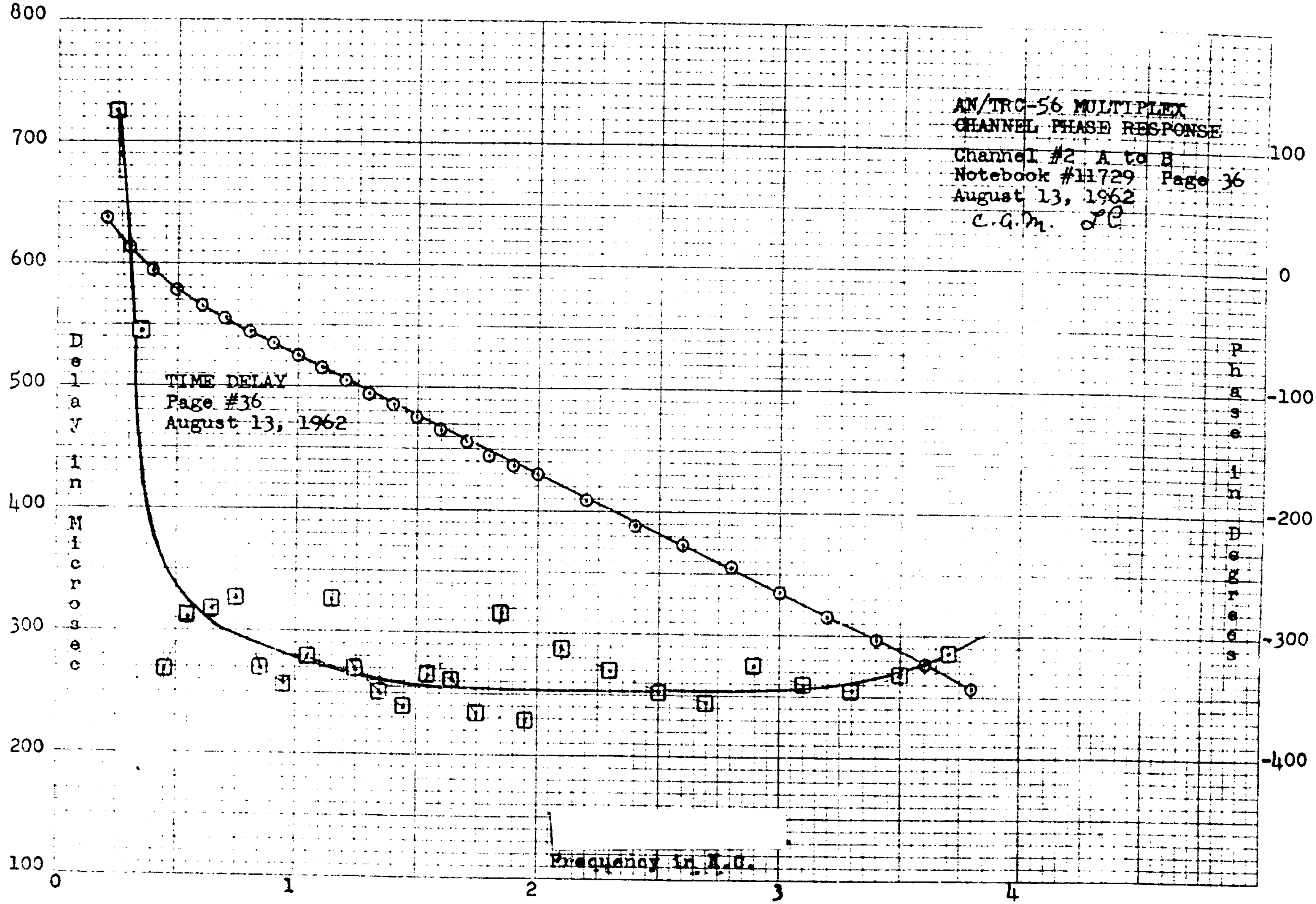


Figure 32 AN/TRC-56 Multiplex Channel Phase Response, Channel 2, A to B

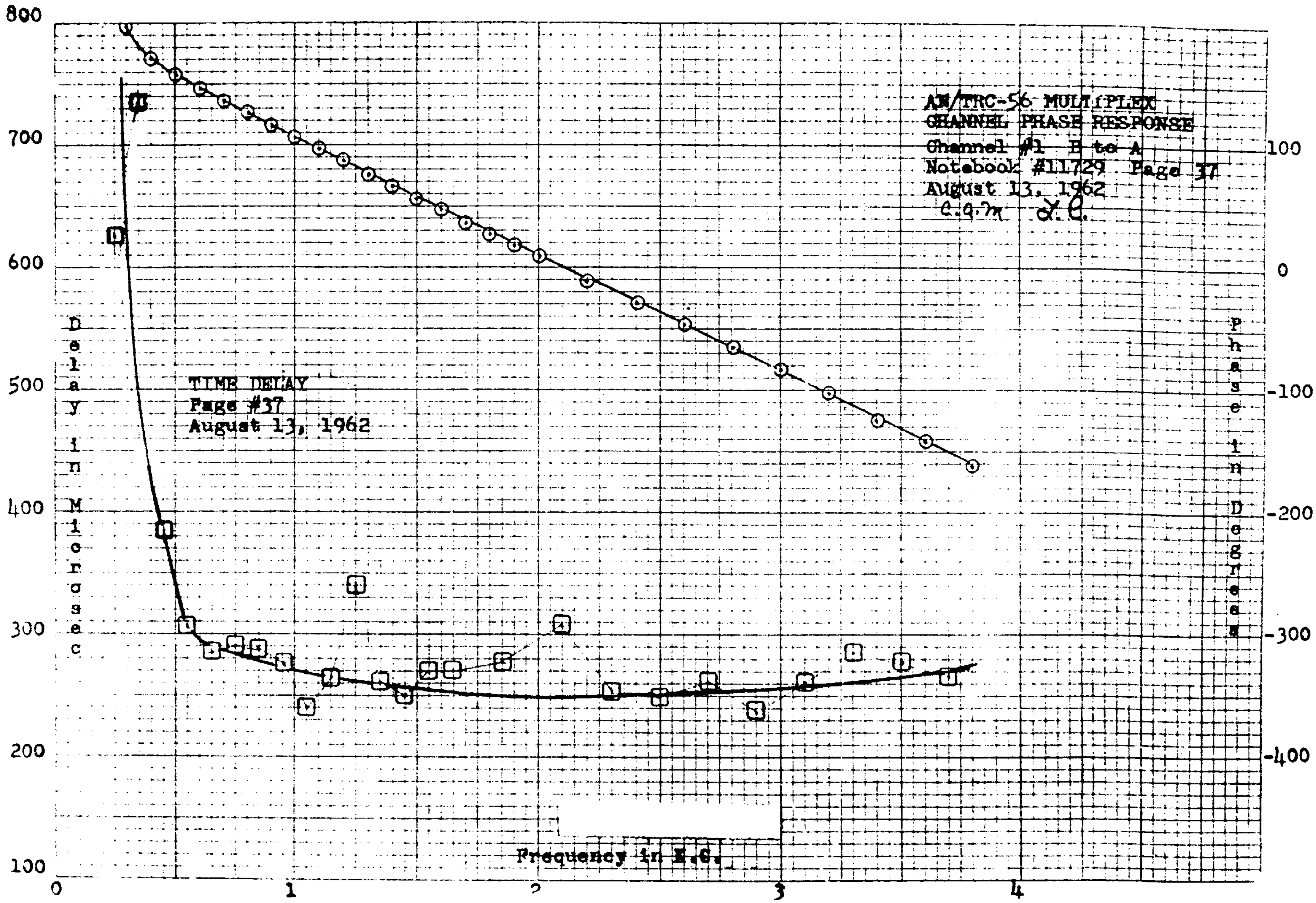


Figure 33 AN/TRC-56 Multiplex Channel Phase Response, Channel 1, B to A

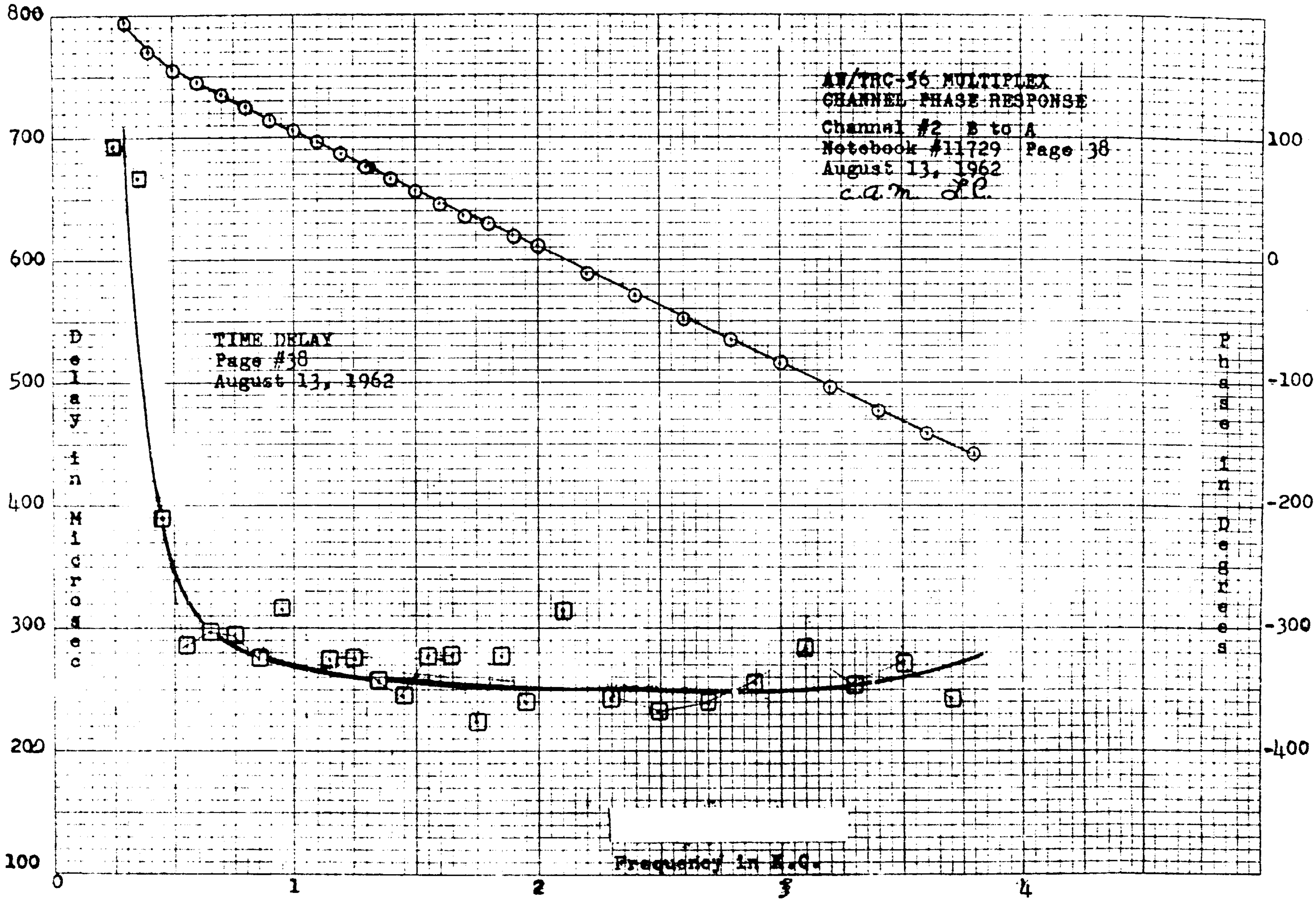
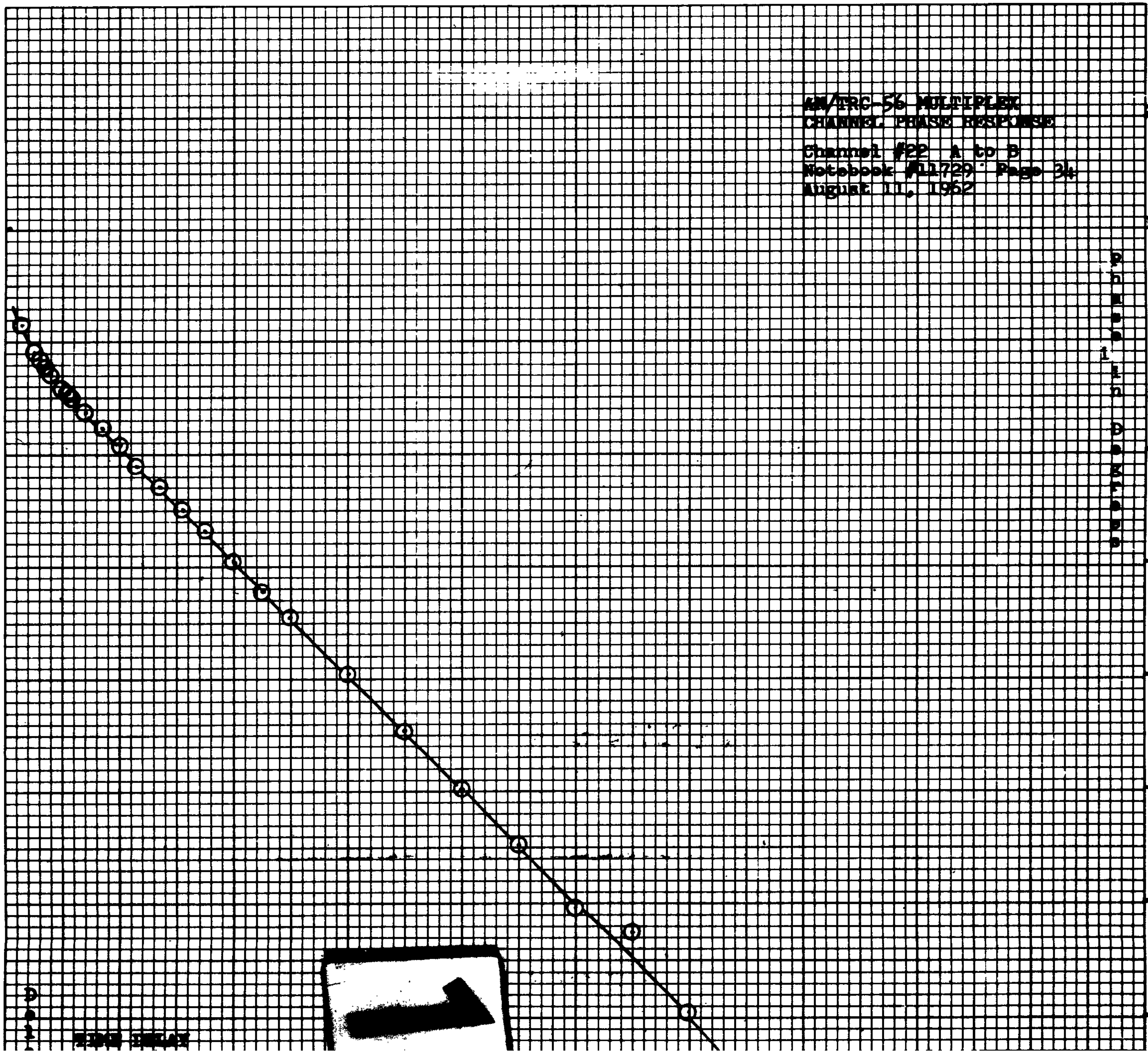


Figure 34 AN/TRC-56 Multiplex Channel Phase Response, Channel 2, B to A



AN/TRC-56 MULTIPLEX
CHANNEL PHASE RESPONSE
Channel #22 A to B
Notebook #11729 Page 34
August 11, 1962
Phase in Degrees
200
100
0
-100
-200
-300
-400
-500
-600

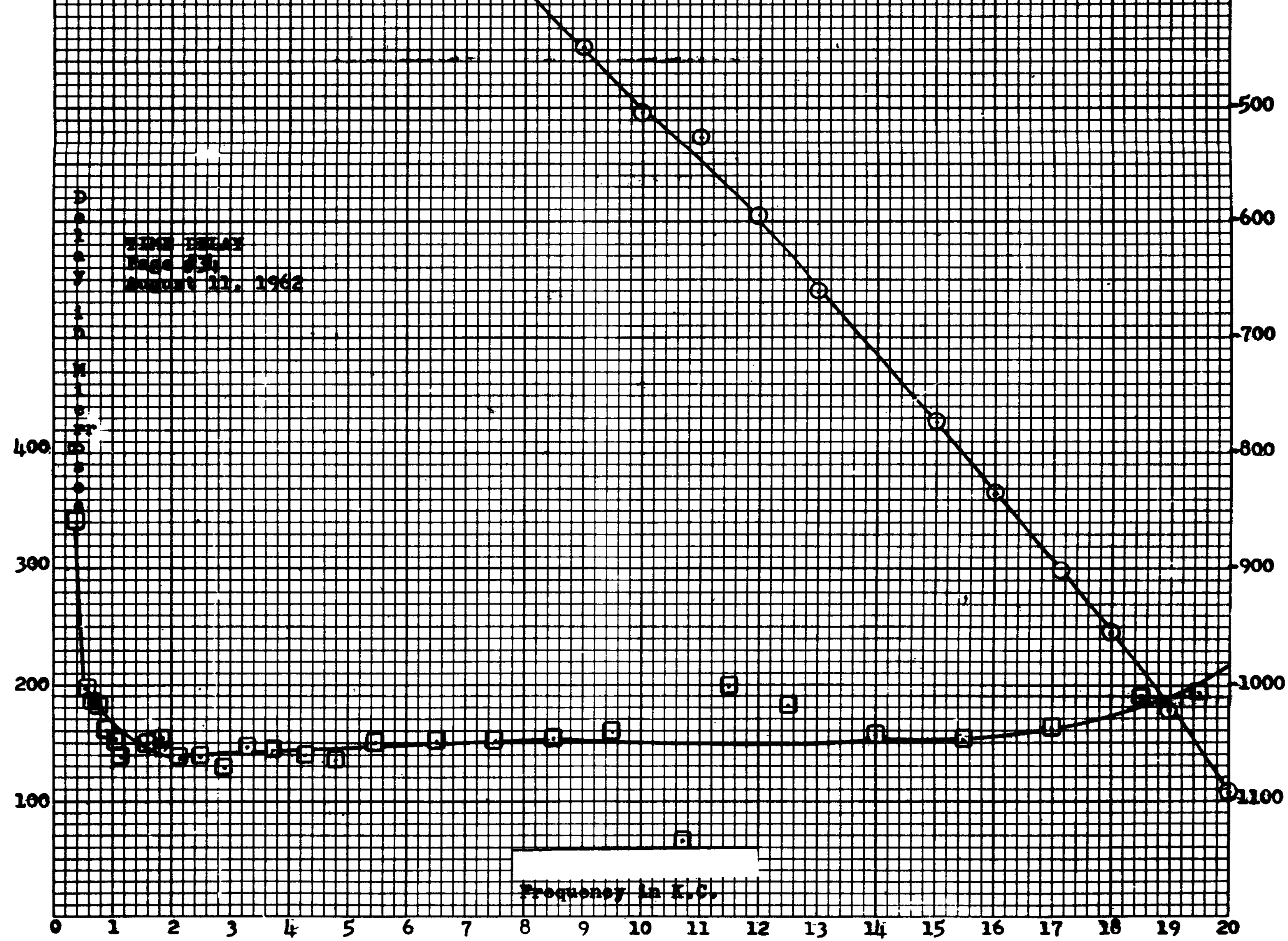
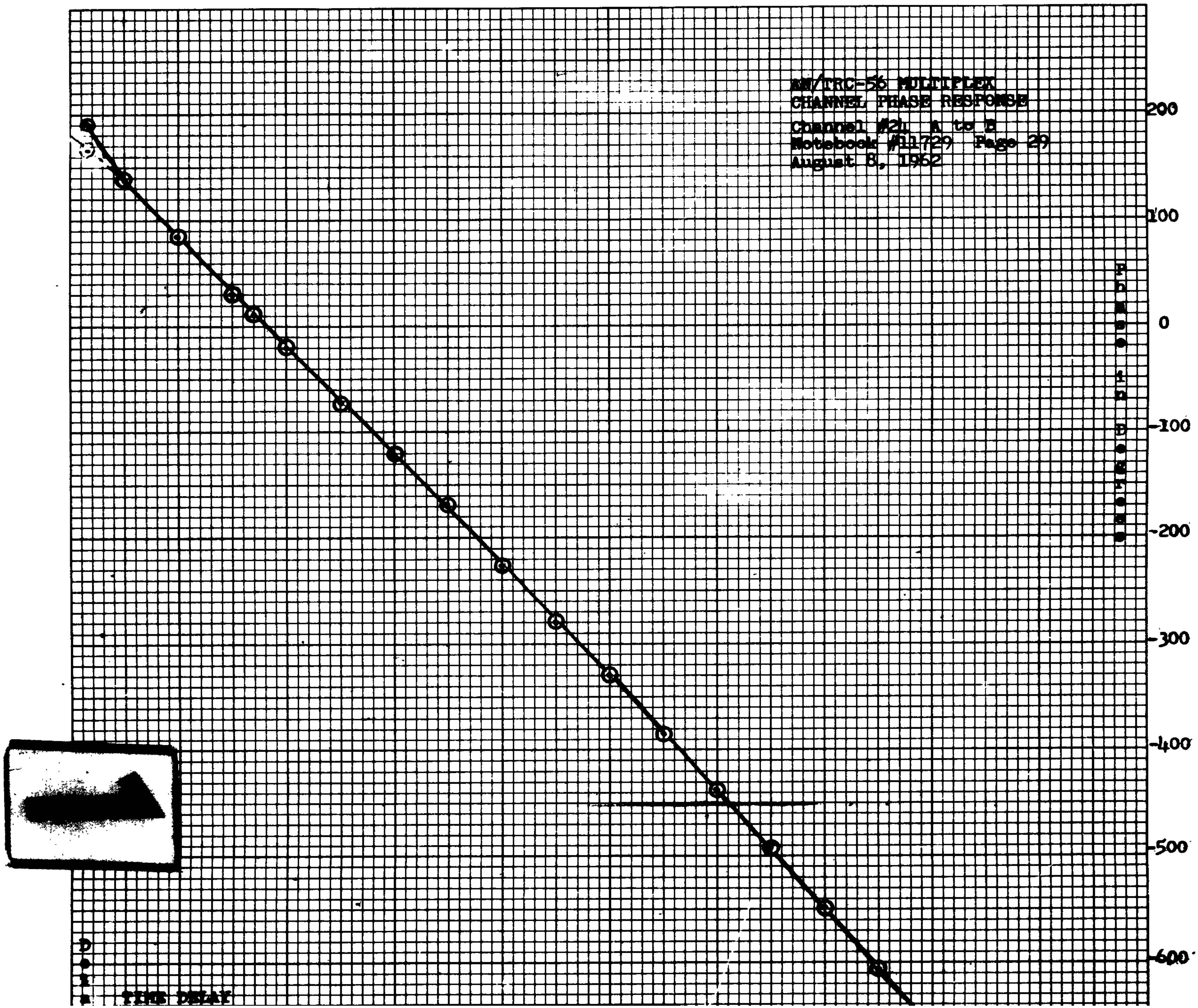


Figure 35 AN/TRC-56 Multiplex Channel Phase Response, Channel 22, A to B


AN/TRC-56 MULTIPLEX
CHANNEL PHASE RESPONSE
Channel #24 A to B
Notebook #11729 Page 29
August 8, 1962
Phase in Degrees
200
100
0
-100
-200
-300
-400
-500
-600
TIME DELAY
1

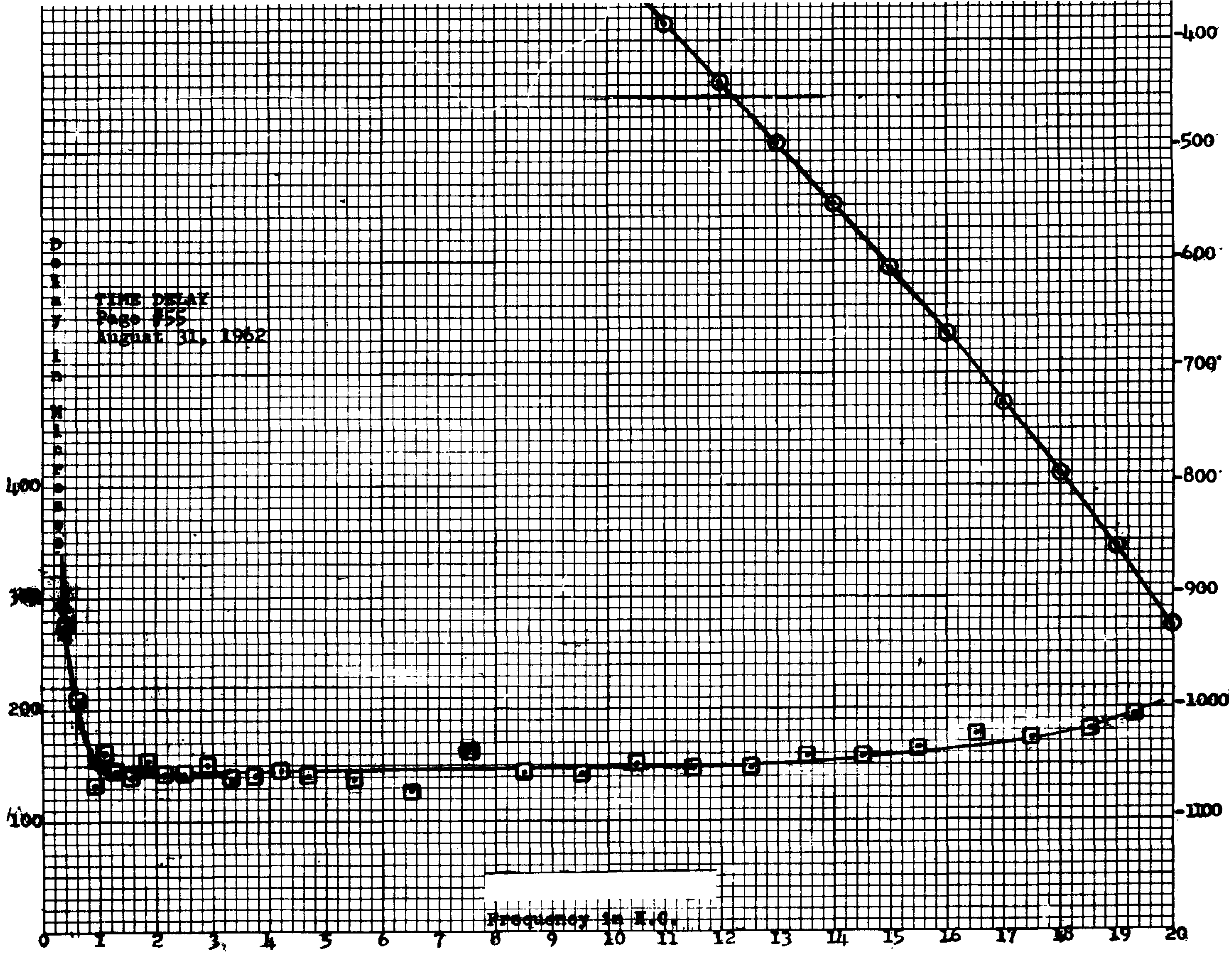
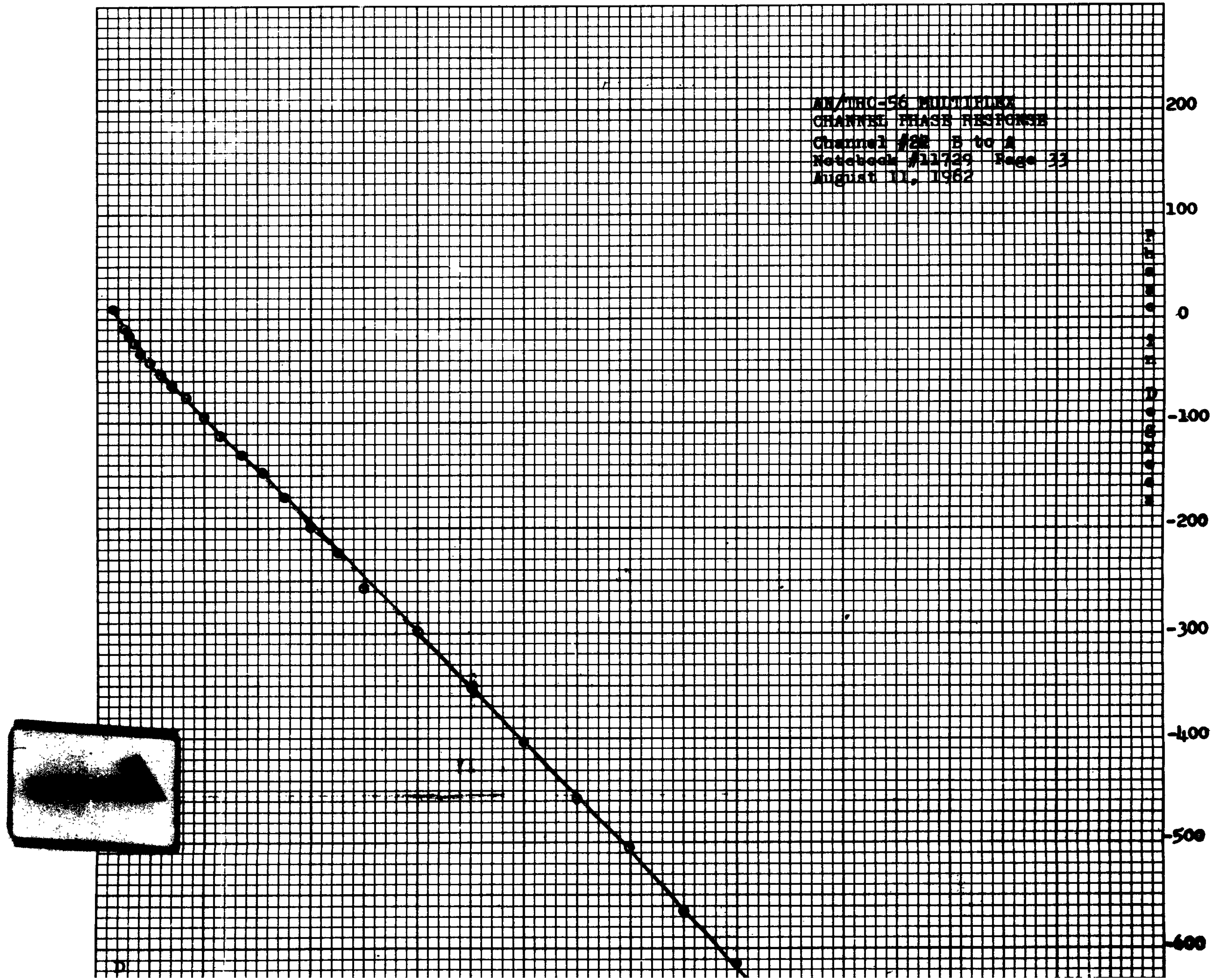


Figure 36 AN/TRC-56 Multiplex Channel Phase Response, Channel 24, A to B


AN/TRC-56 MULTIPLEX
CHANNEL PHASE RESPONSE
Channel #22 B to A
Notebook #11729 Page 33
August 11, 1962
Phase in Degrees
200
100
0
-100
-200
-300
-400
-500
-600

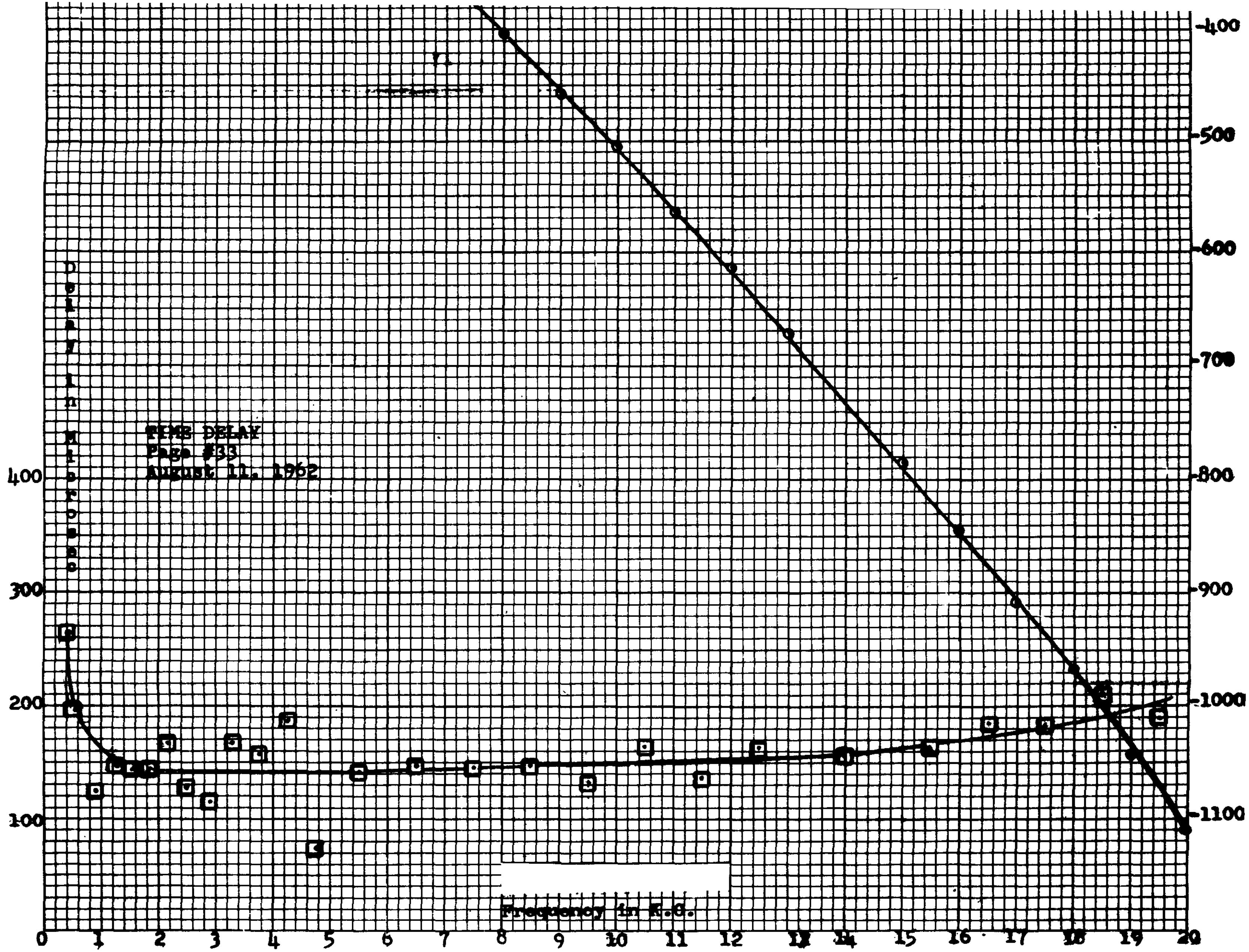
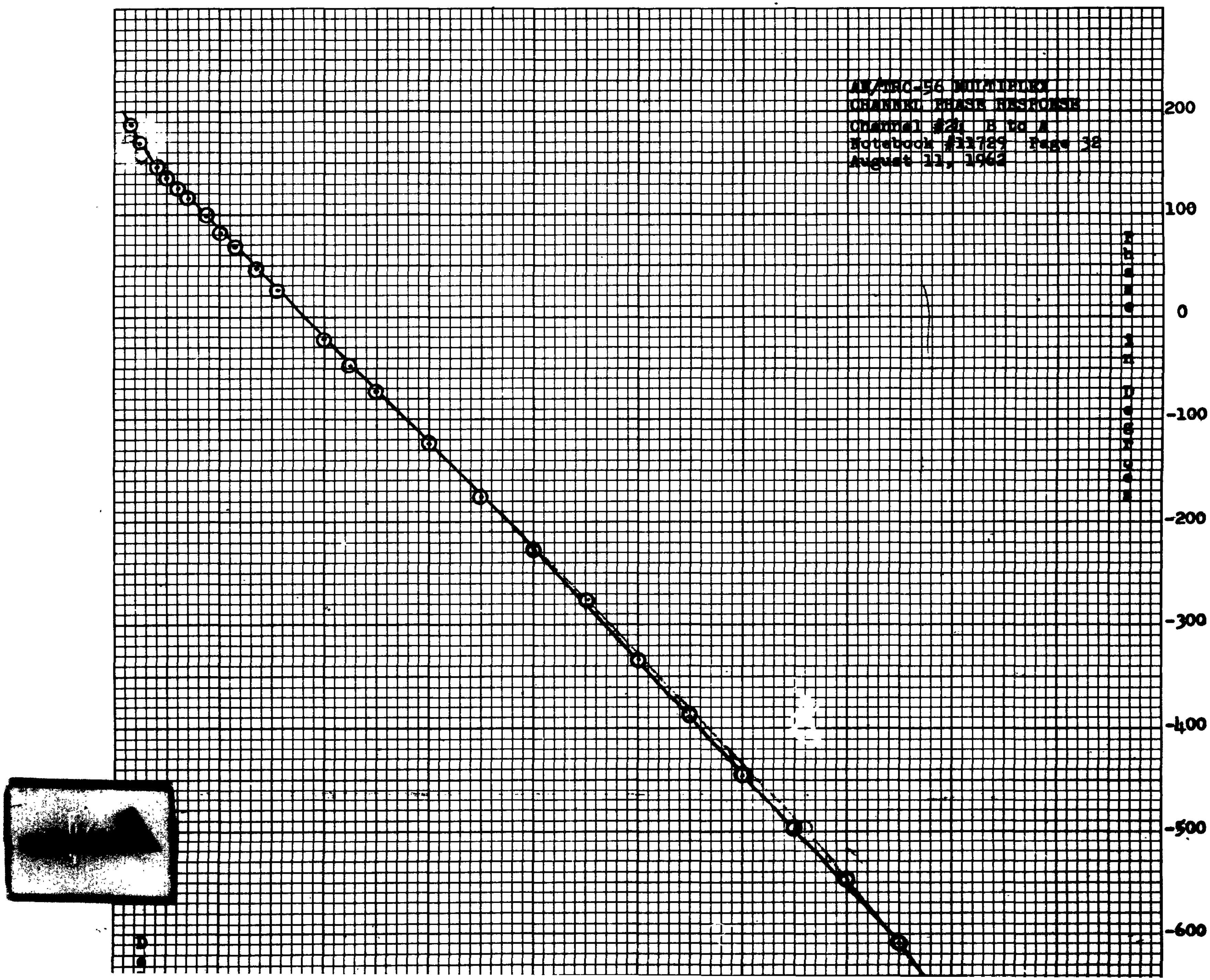


Figure 37 AN/TRC-56 Multiplex Channel Phase Response, Channel 22, B to A


AN/TRC-56 MULTIPLEX
CHANNEL PHASE RESPONSE
Channel #21 B to A
Notebook #11729 Page 32
August 11, 1962
Phase in Degrees
200
100
0
-100
-200
-300
-400
-500
-600

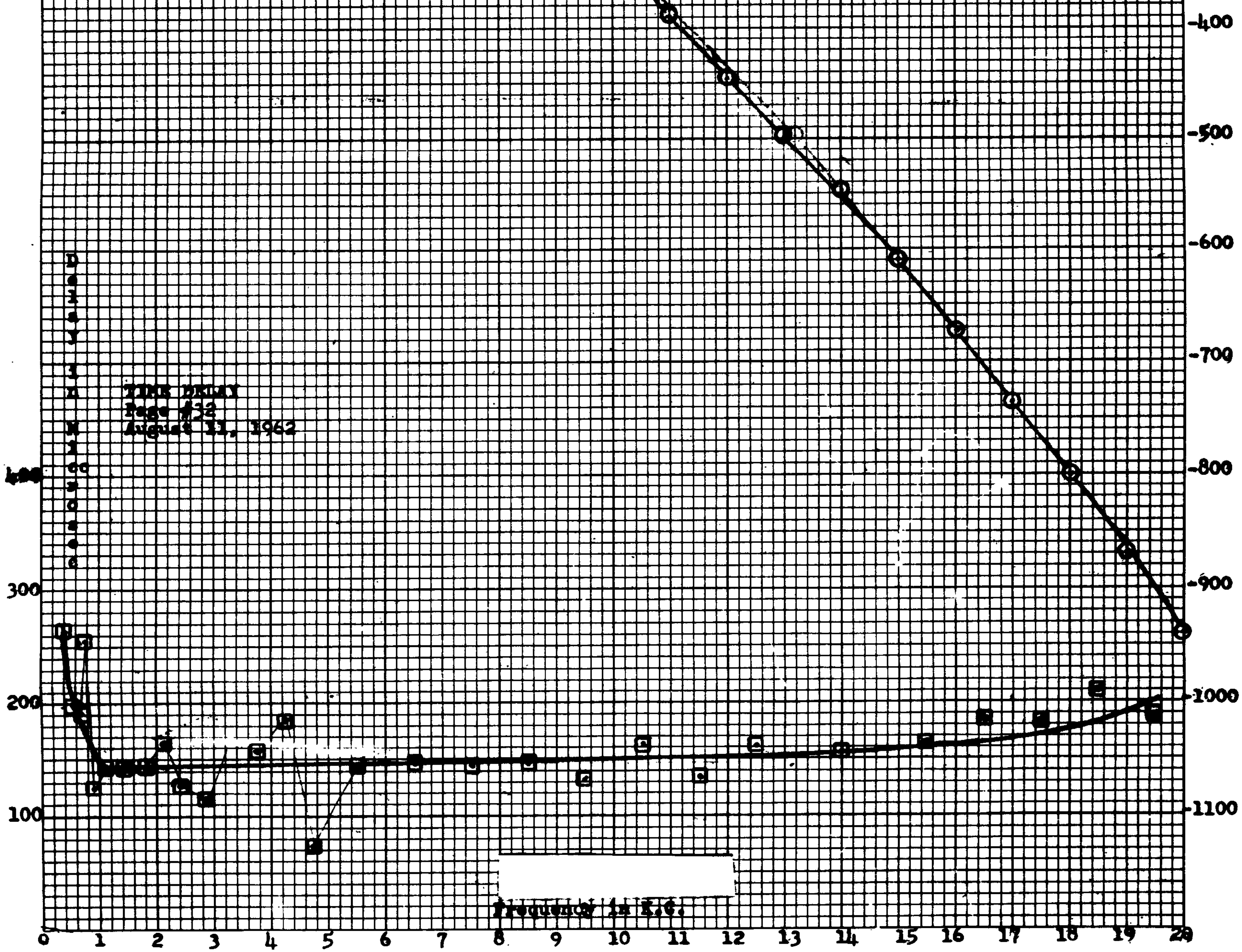


Figure 38 AN/TRC-56 Multiplex Channel Phase Response, Channel 24, B to A

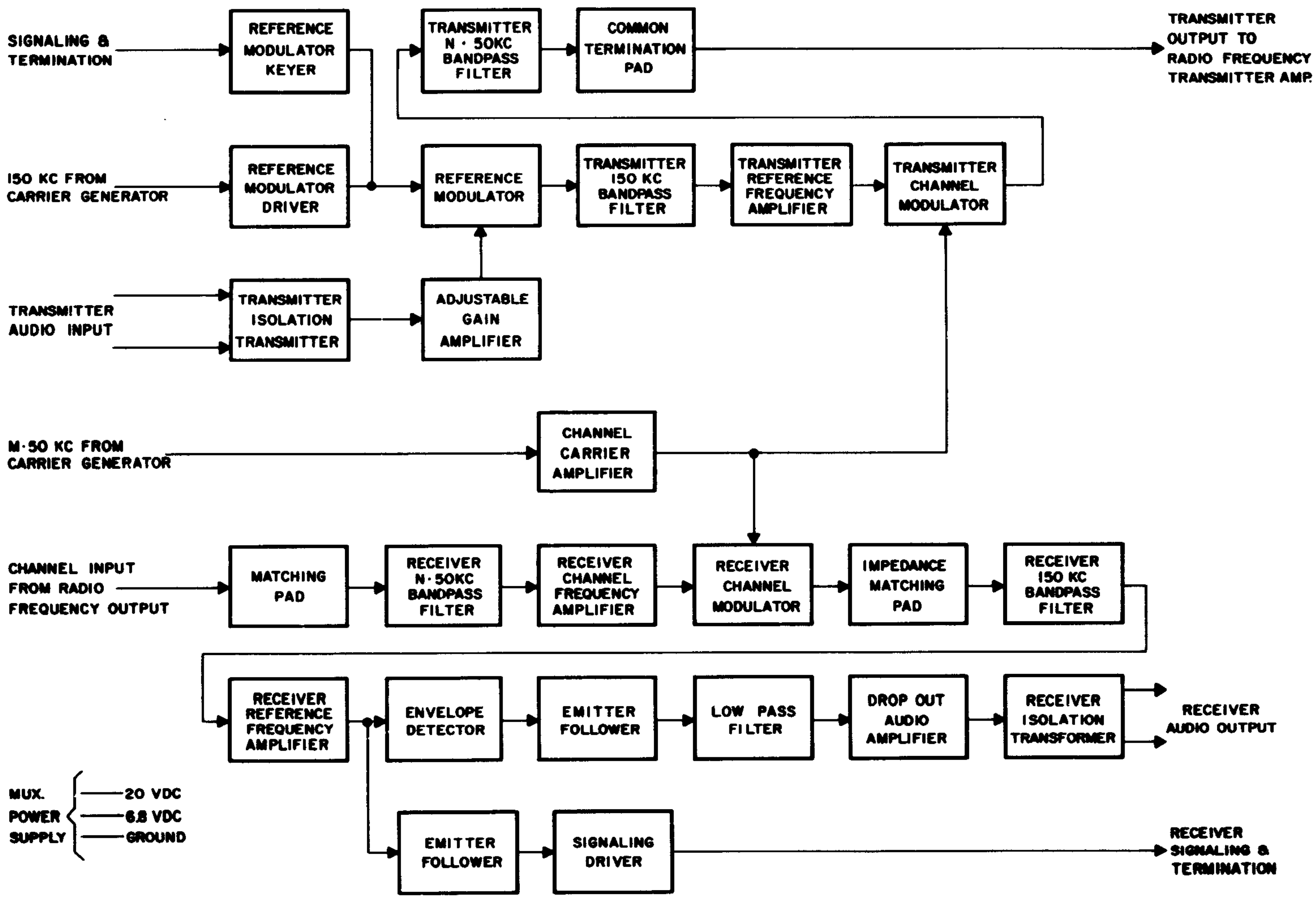
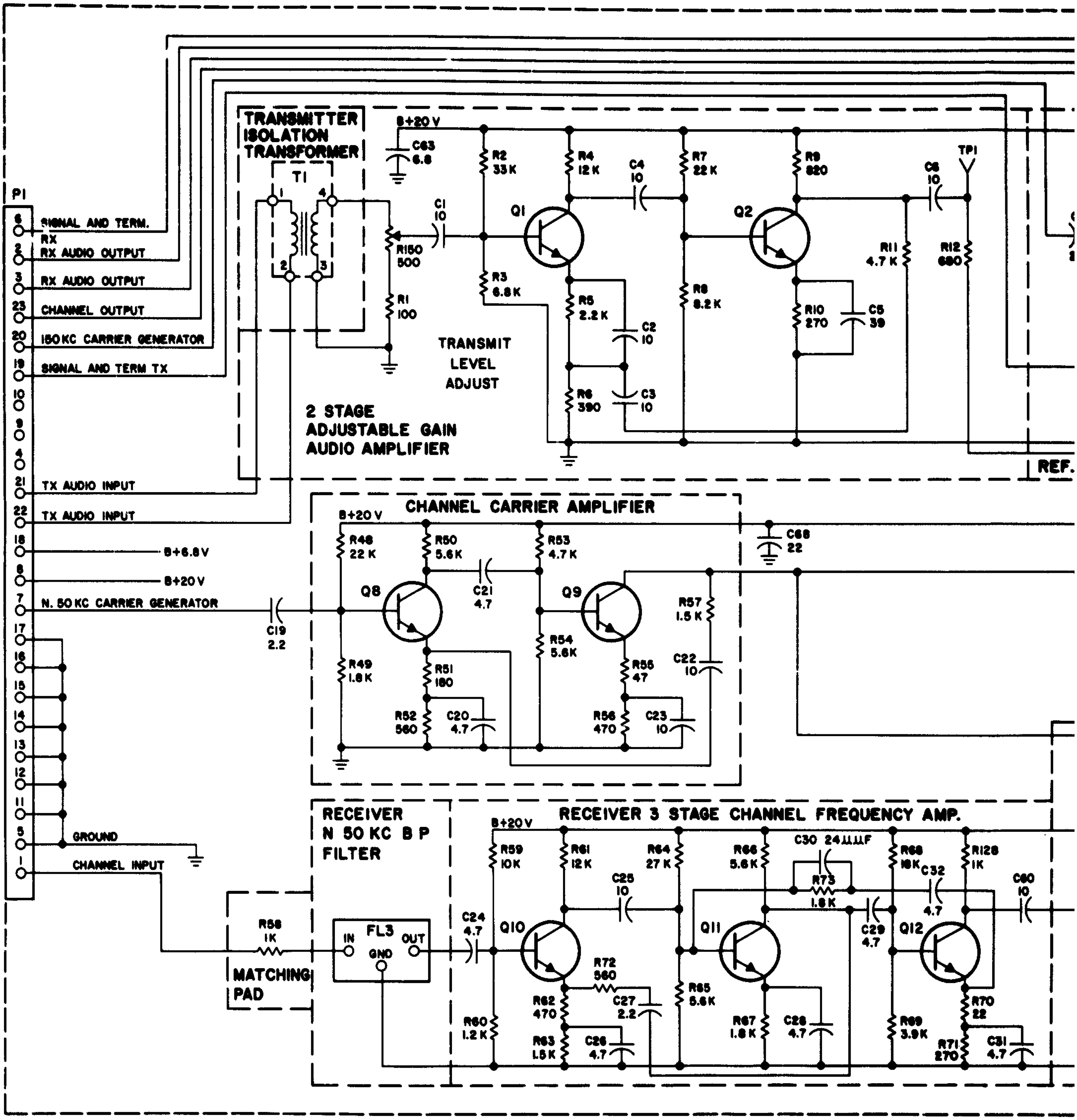


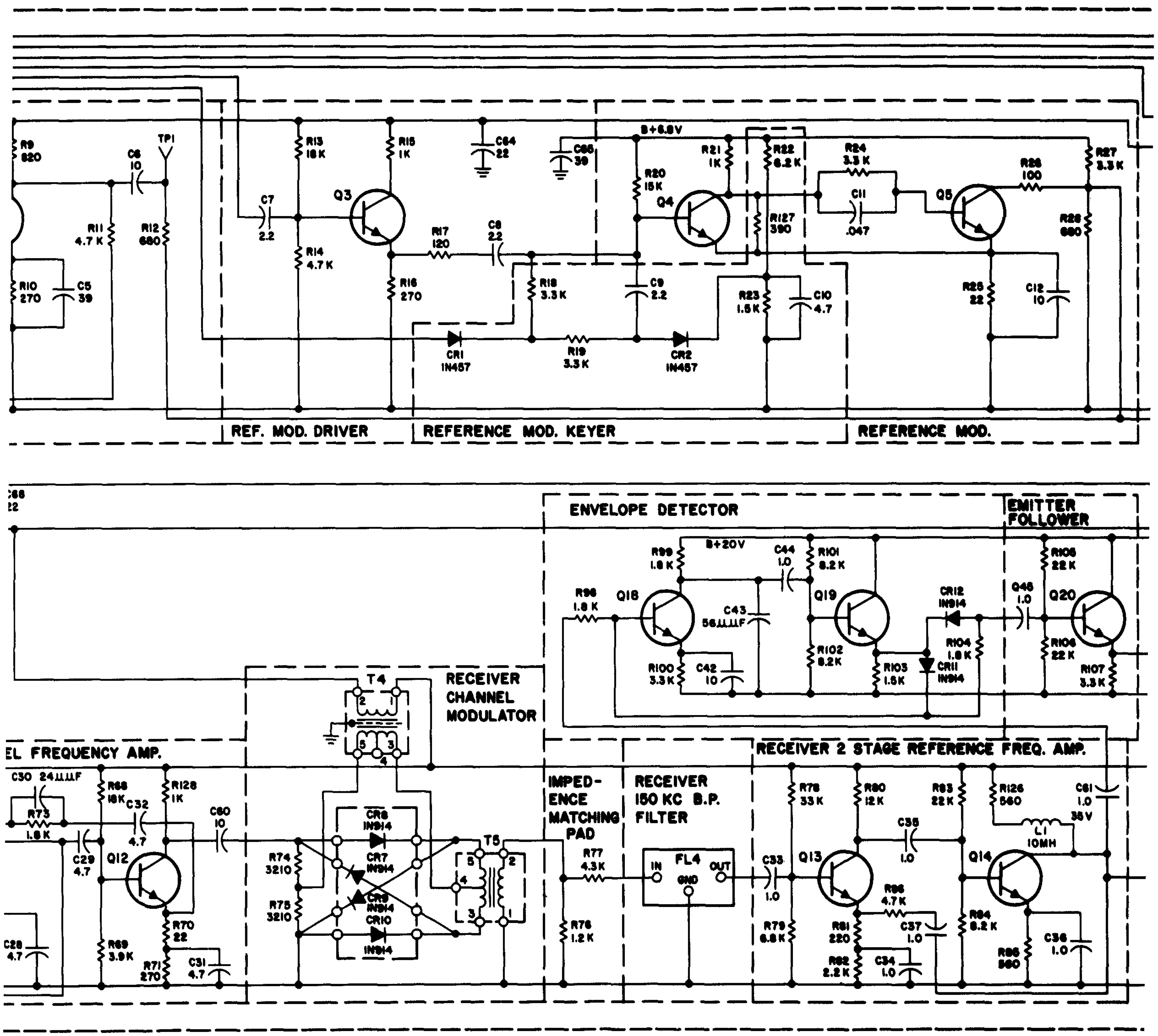
Figure 39 Transmit-Receive Assembly, Block Diagram



P1
6 SIGNAL AND TERM. RX
2 RX AUDIO OUTPUT
3 RX AUDIO OUTPUT
23 CHANNEL OUTPUT
20 150KC CARRIER GENERATOR
19 SIGNAL AND TERM TX
10
9
4
21 TX AUDIO INPUT
22 TX AUDIO INPUT
18 B+6.8V
8 B+20V
7 N. 50KC CARRIER GENERATOR
17
16
15
14
13
12
11
5 GROUND
1 CHANNEL INPUT
TRANSMITTER ISOLATION TRANSFORMER
T1
B+20V
C63 6.8
R2 33K
R4 12K
C4 10
R7 22K
R9 820
C6 10
TP1
C1 10
R150 500
Q1
Q2
R11 4.7K
R12 680
R1 100
R3 6.8K
R5 2.2K
C2 10
R8 8.2K
R10 270
C5 39
TRANSMIT LEVEL ADJUST
R6 390
C3 10
2 STAGE ADJUSTABLE GAIN AUDIO AMPLIFIER
REF.
CHANNEL CARRIER AMPLIFIER
B+20V
R48 22K
R50 5.6K
R53 4.7K
C68 22
Q8
C21 4.7
Q9
R57 1.5K
C19 2.2
R54 5.6K
R49 1.8K
R51 180
R55 47
C22 10
R52 560
C20 4.7
R56 470
C23 10
RECEIVER N 50 KC B P FILTER
RECEIVER 3 STAGE CHANNEL FREQUENCY AMP.
B+20V
R59 10K
R61 12K
R64 27K
R66 5.6K
C30 24µµF
R68 18K
R128 1K
C25 10
R73 1.8K
C32 4.7
C60 10
C24 4.7
Q10
Q11
C29 4.7
Q12
R58 1K
FL3
IN
GND
OUT
MATCHING PAD
R72 560
R62 470
C27 2.2
R65 5.6K
R67 1.8K
C28 4.7
R70 22
R60 1.2K
R63 1.5K
C26 4.7
R69 3.9K
R71 270
C31 4.7

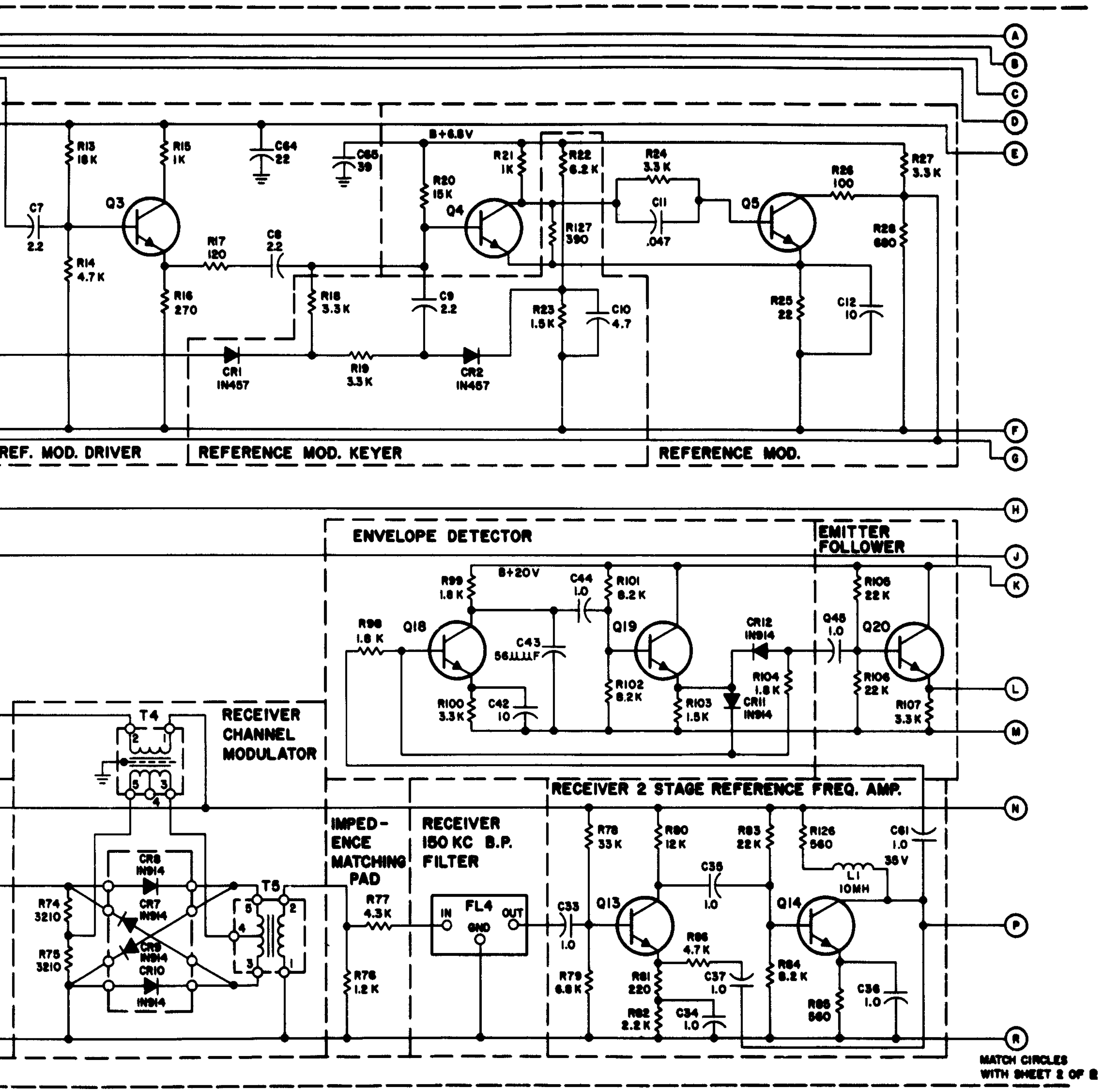
1

SHEE



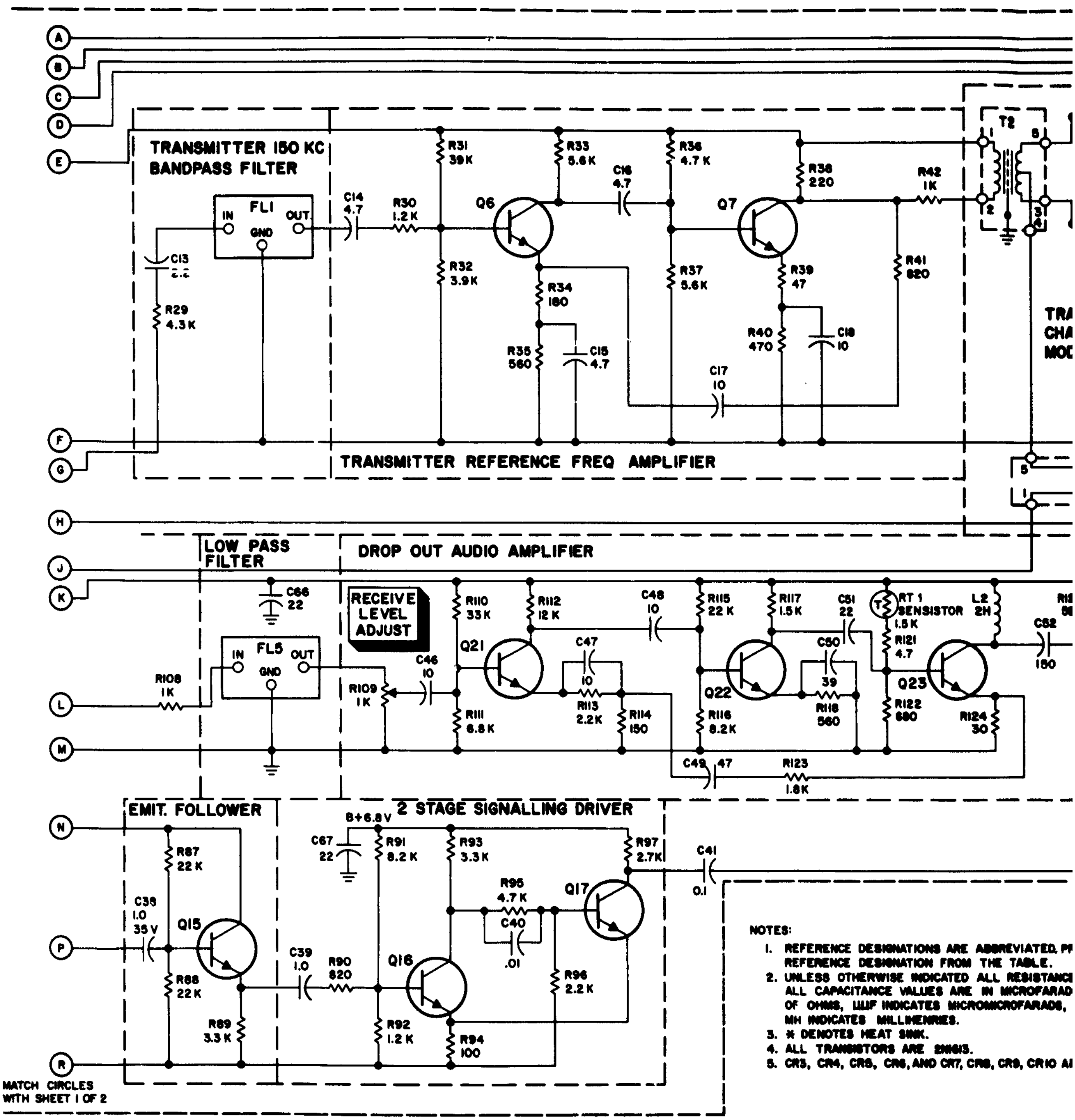
SHEET 1 OF 2

Figure 40 Transmit-Receive Assembly, Schema



SHEET 1 OF 2

Figure 40 Transmit-Receive Assembly, Schematic Diagram

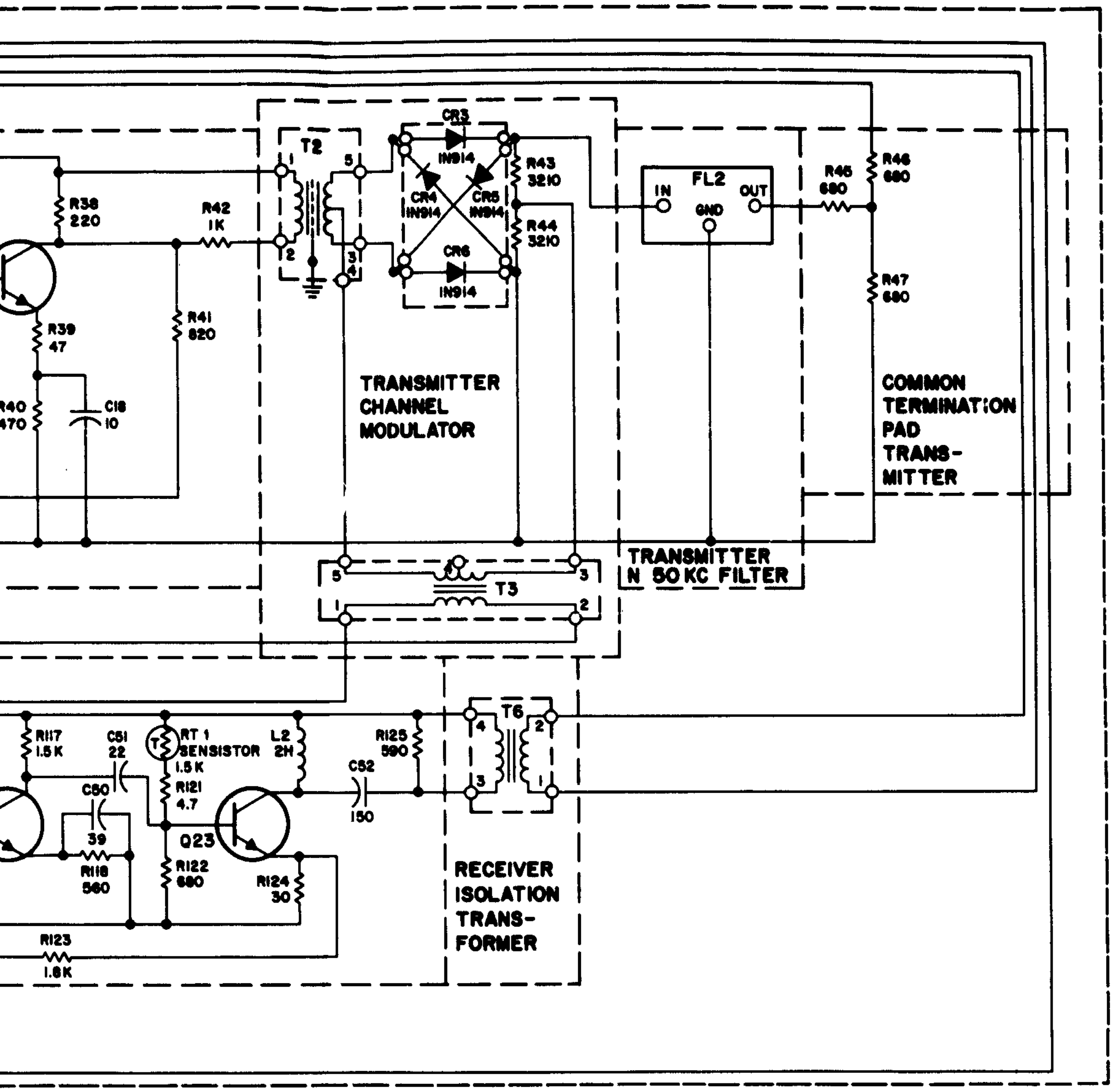


NOTES:

1. REFERENCE DESIGNATIONS ARE ABBREVIATED. PF
REFERENCE DESIGNATION FROM THE TABLE.
2. UNLESS OTHERWISE INDICATED ALL RESISTANCE
ALL CAPACITANCE VALUES ARE IN MICROFARAD
OF OHMS, UUF INDICATES MICROMICROFARADS,
MH INDICATES MILLIHENRIES.
3. * DENOTES HEAT SINK.
4. ALL TRANSISTORS ARE 2N613.
5. CR3, CR4, CR5, CR6, AND CR7, CR8, CR9, CR10 A

1

SHEET 2



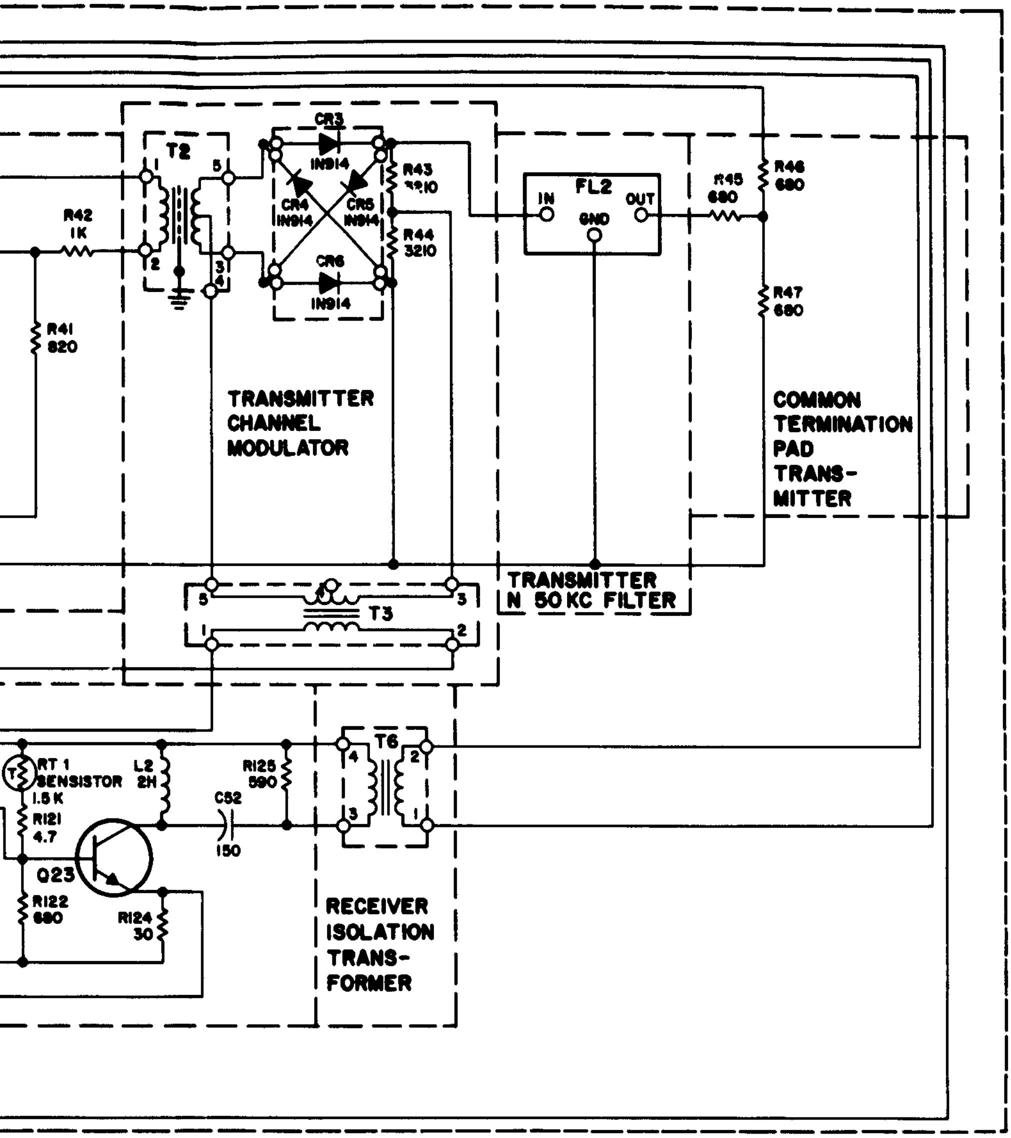
NOTES:

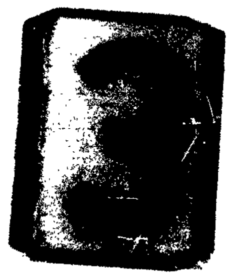
1. REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE APPROPRIATE REFERENCE DESIGNATION FROM THE TABLE.
2. UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS, ALL CAPACITANCE VALUES ARE IN MICROFARADS. K INDICATES THOUSANDS OF OHMS, UUF INDICATES MICROMICROFARADS, H INDICATES HENRIES, MH INDICATES MILLIHENRIES.
3. * DENOTES HEAT SINK.
4. ALL TRANSISTORS ARE 2N1613.
5. CR3, CR4, CR5, CR6, AND CR7, CR8, CR9, CR10 ARE IN914 DIODES IN MATCHED QUADS.

	VOICE CHANNELS												DATA CHANNELS							
CHANNEL	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20
FREQUENCY	650 KC	950 KC	700 KC	1.0MC	750 KC	1050 KC	800 KC	1100 KC	850 KC	1150 KC	900 KC	1.2 MC	50KC	350 KC	300 KC	400 KC	150KC	450 KC	200 KC	500 KC
MIL. REF. DESIGNATION	1A1	1A2	1A3	1A4	1A5	1A6	1A7	1A8	1A9	1A10	1A11	1A12	1A13	1A14	1A15	1A16	1A17	1A18	1A19	1A20
PHILCO ASSEMBLY NUMBERS	340-[illegible]-1	340-[illegible]-2	340-[illegible]-3	340-[illegible]-4	340-[illegible]-5	340-[illegible]-6	340-[illegible]-7	340-[illegible]-8	340-[illegible]-9	340-[illegible]-10	340-[illegible]-11	340-[illegible]-12	340-[illegible]-13	340-[illegible]-14	[illegible]	[illegible]	340-[illegible]-17	[illegible]	[illegible]	[illegible]

SHEET 2 OF 2

Figure 40 Transmit-Receive Assembly, Schemat





	VOICE CHANNELS												DATA CHANNELS											
CHANNEL	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24
FREQUENCY	650 KC	950 KC	700 KC	1.0MC	750 KC	1050 KC	800 KC	1100 KC	850 KC	1150 KC	900 KC	1.2MC	50KC	350 KC	100 KC	400 KC	150KC	450 KC	200 KC	500 KC	250 KC	550 KC	300 KC	600 KC
MIL. REF. DESIGNATION	1A1	1A2	1A3	1A4	1A5	1A6	1A7	1A8	1A9	1A10	1A11	1A12	1A13	1A14	1A15	1A16	1A17	1A18	1A19	1A20	1A21	1A22	1A23	1A24
PHILCO ASSEMBLY NUMBERS	340-5868 1	340-5868 2	340-5868 3	340-5868 4	340-5868 5	340-5868 6	340-5868 7	340-5868 8	340-5868 9	340-5868 10	340-5868 11	340-5868 12	340-5868 13	340-5868 14	340-5868 15	340-5868 16	340-5868 17	340-5868 18	340-5868 19	340-5868 20	340-5868 21	340-5868 22	340-5868 23	340-5868 24

ESIGNATIONS ARE ABBREVIATED, PREFIX THE APPROPRIATE
ESIGNATION FROM THE TABLE.
RWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS,
NCE VALUES ARE IN MICROFARADS. K INDICATES THOUSANDS
F INDICATES MICROMICROFARADS, H INDICATES HENRIES,
MILLIHENRIES.
EAT SINK.
TORS ARE 2N613.
, CR6, AND CR7, CR8, CR9, CR10 ARE IN914 DIODES IN MATCHED QUADS.

Figure 40 Transmit-Receive Assembly, Schematic Diagram

SHEET 2 OF 2

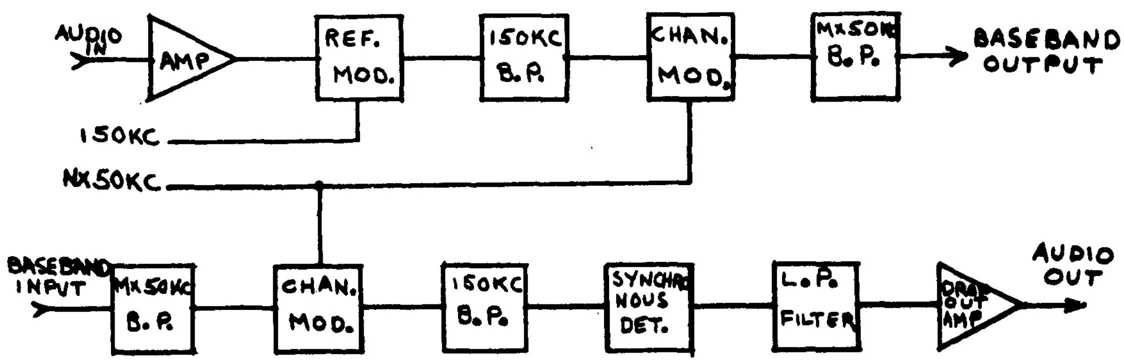


Figure 41 Transmit-Receive Functions, Simplified Block Diagram

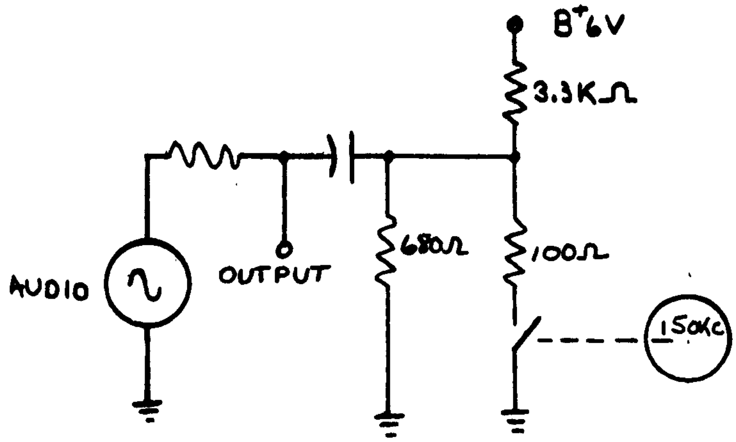


Figure 42 Reference Modulator, Analogous Circuit

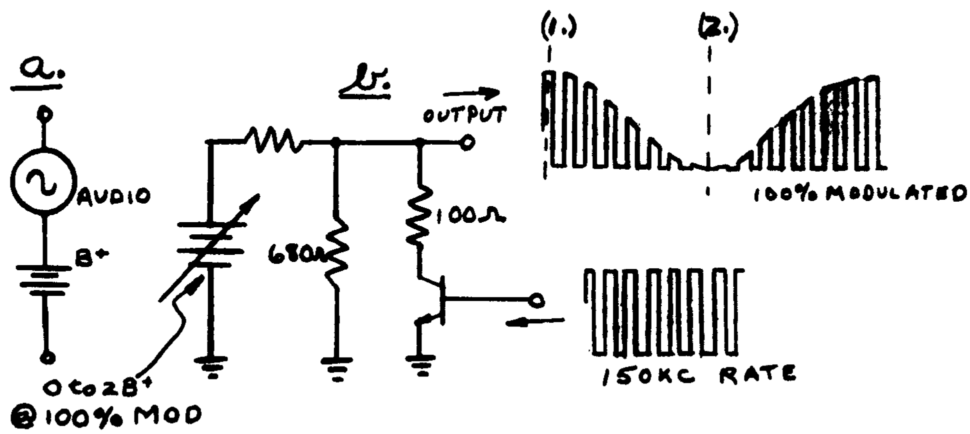


Figure 43 Reference Modulator, Simplified Schematic Diagram

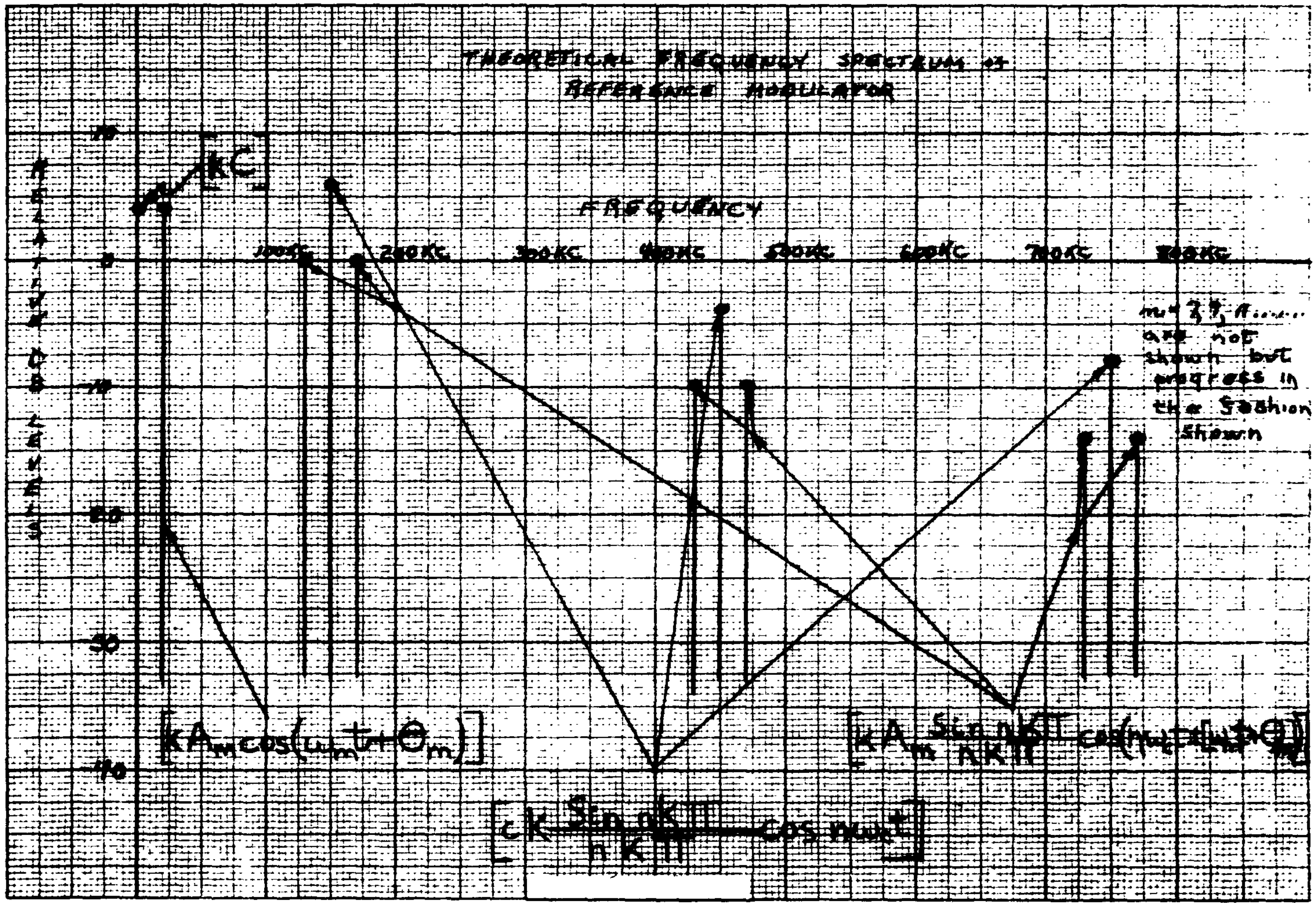


Figure 44 Theoretical Frequency Spectrum of Reference Modulator

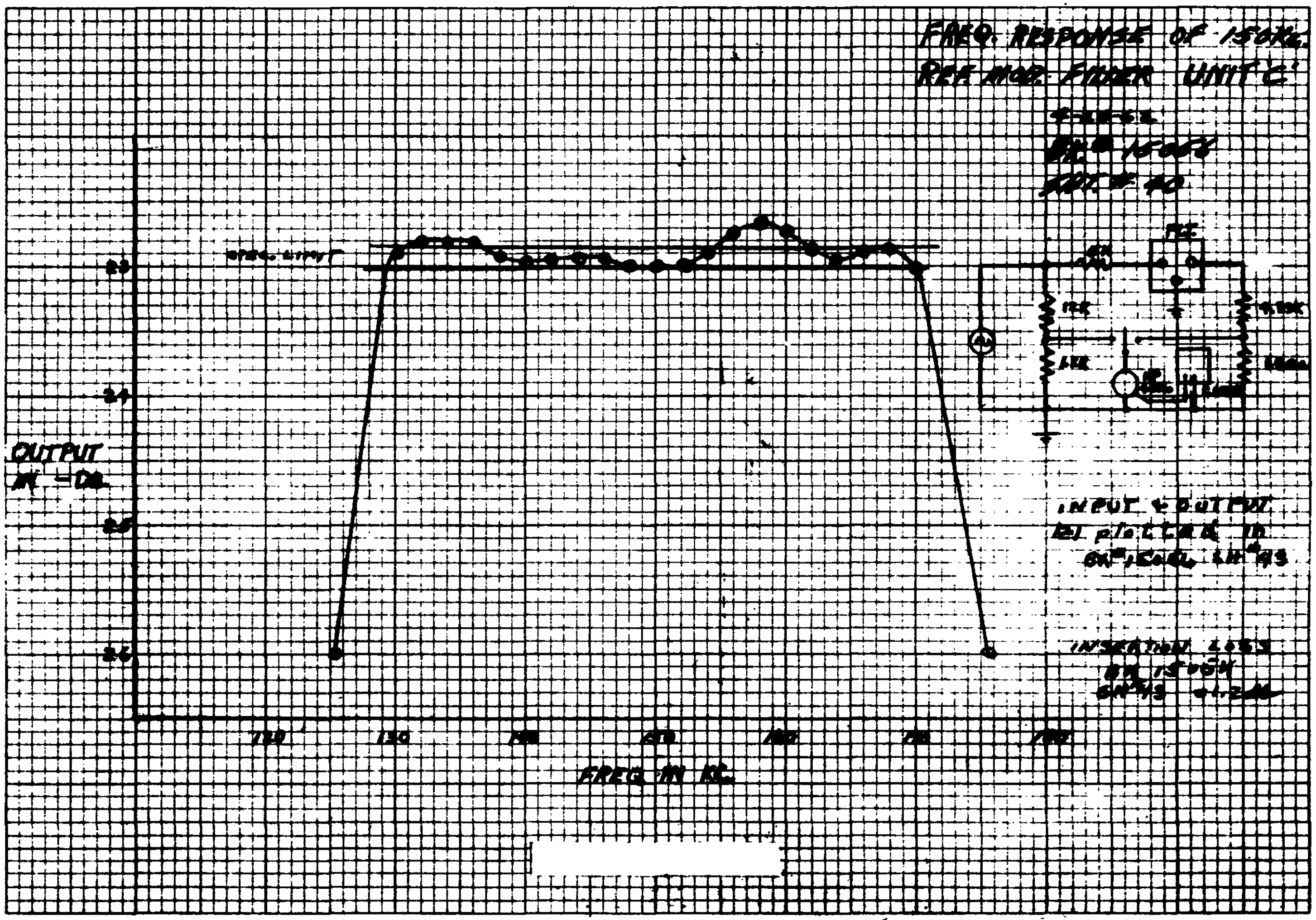


Figure 45 Frequency Response of 150-kc Reference Modulator Filter

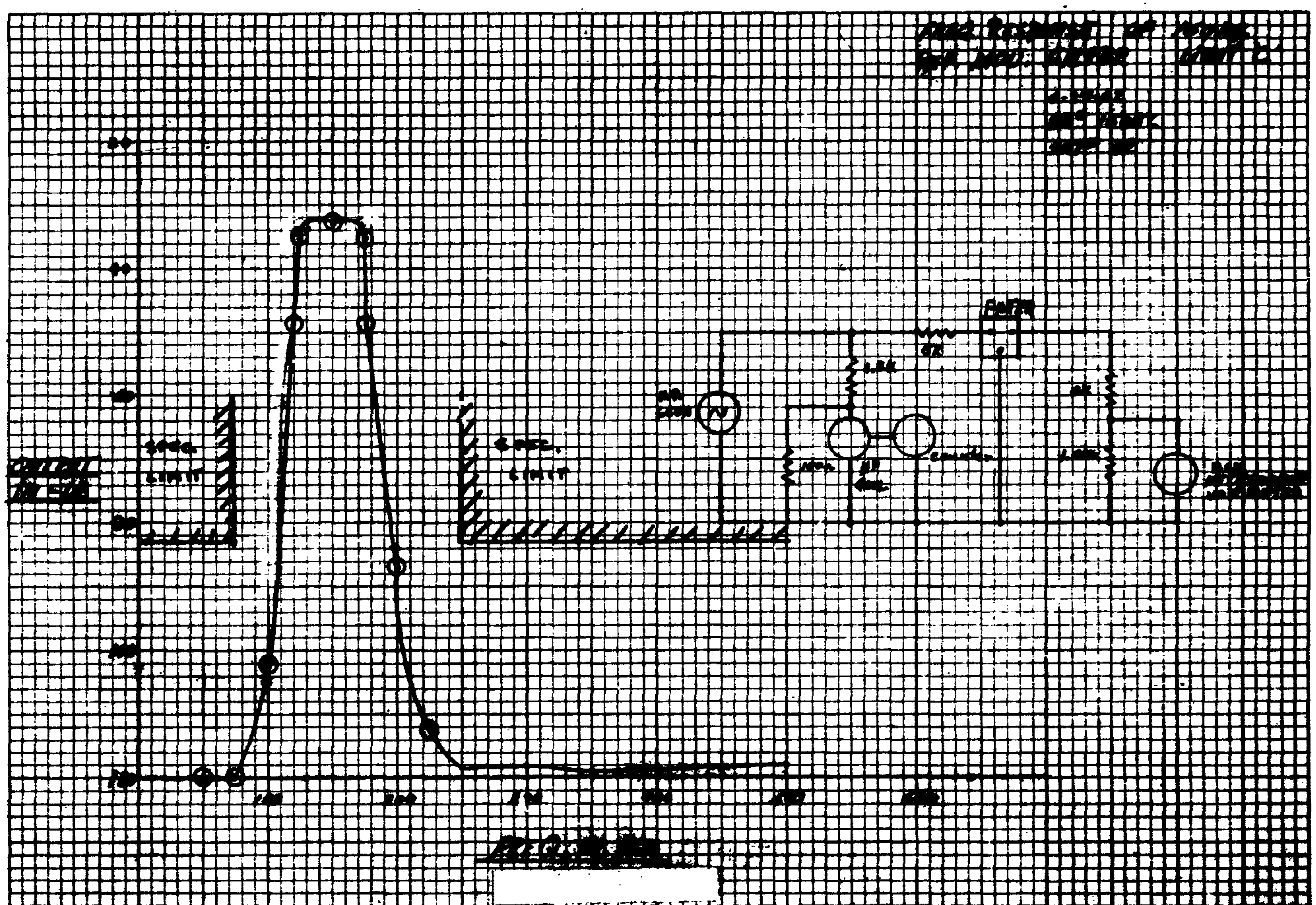


Figure 46 Frequency Response of 150-kc Reference Modulator Filter

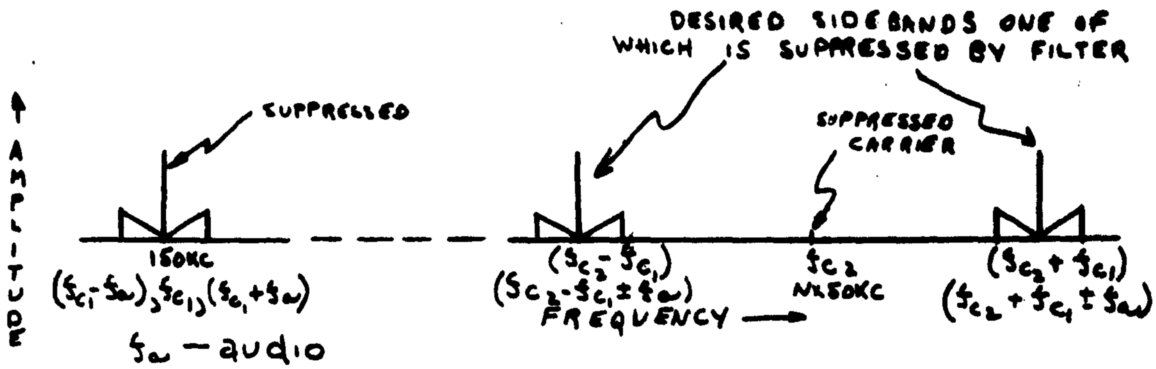


Figure 47 Desired Channel Modulator Output

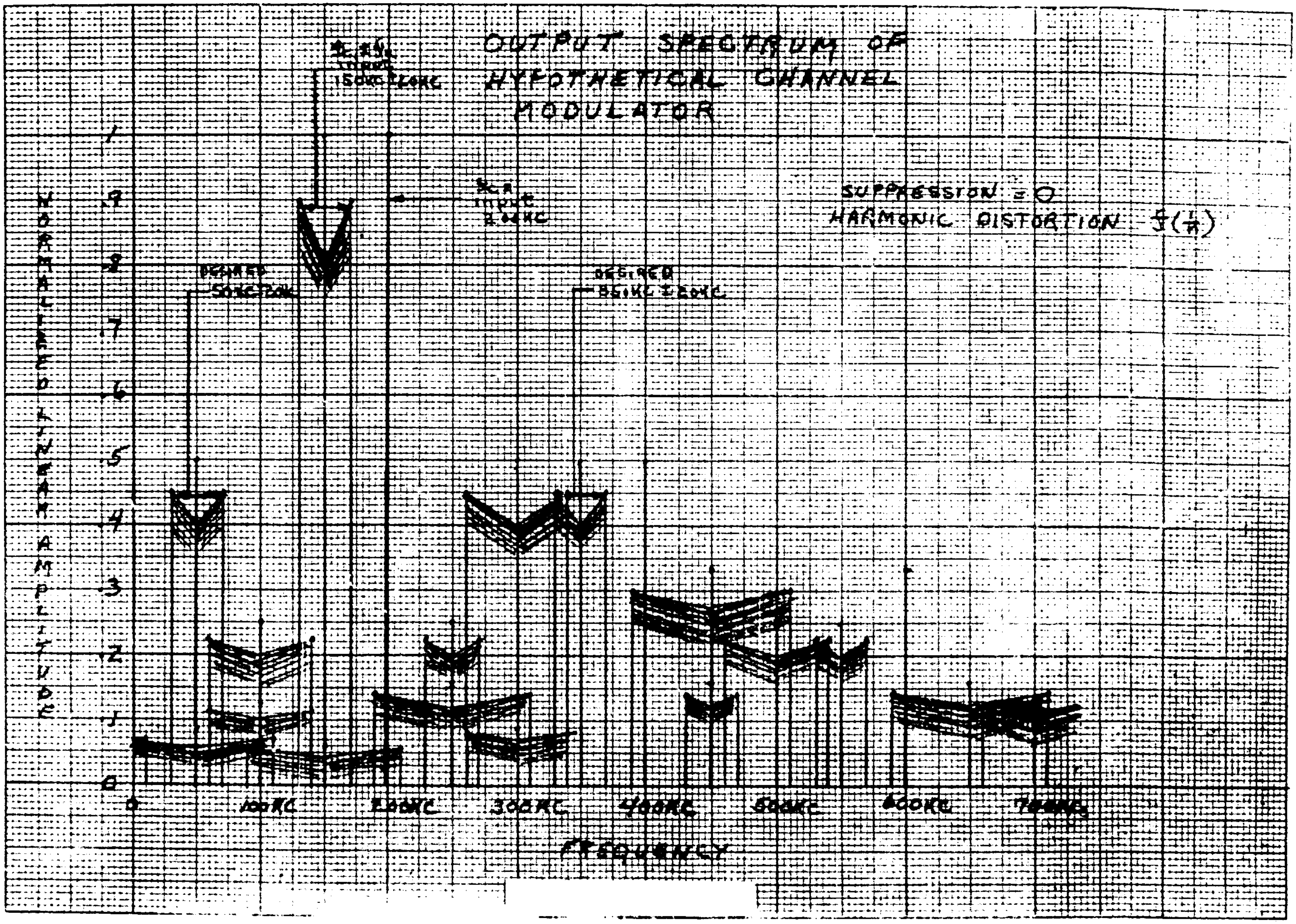


Figure 48 Output Spectrum of Hypothetical Channel Modulator

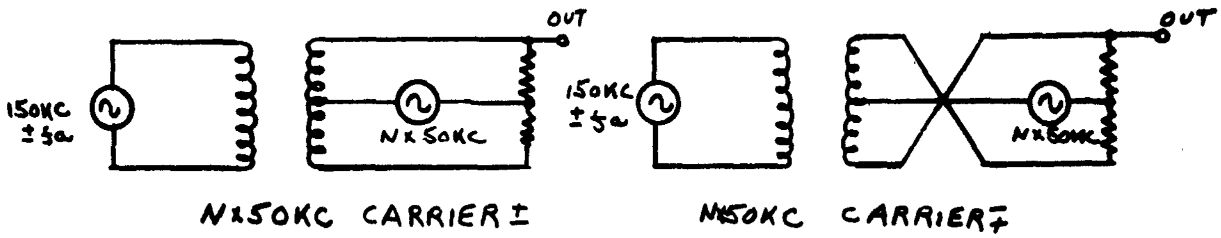


Figure 49 Channel Modulator Switching States

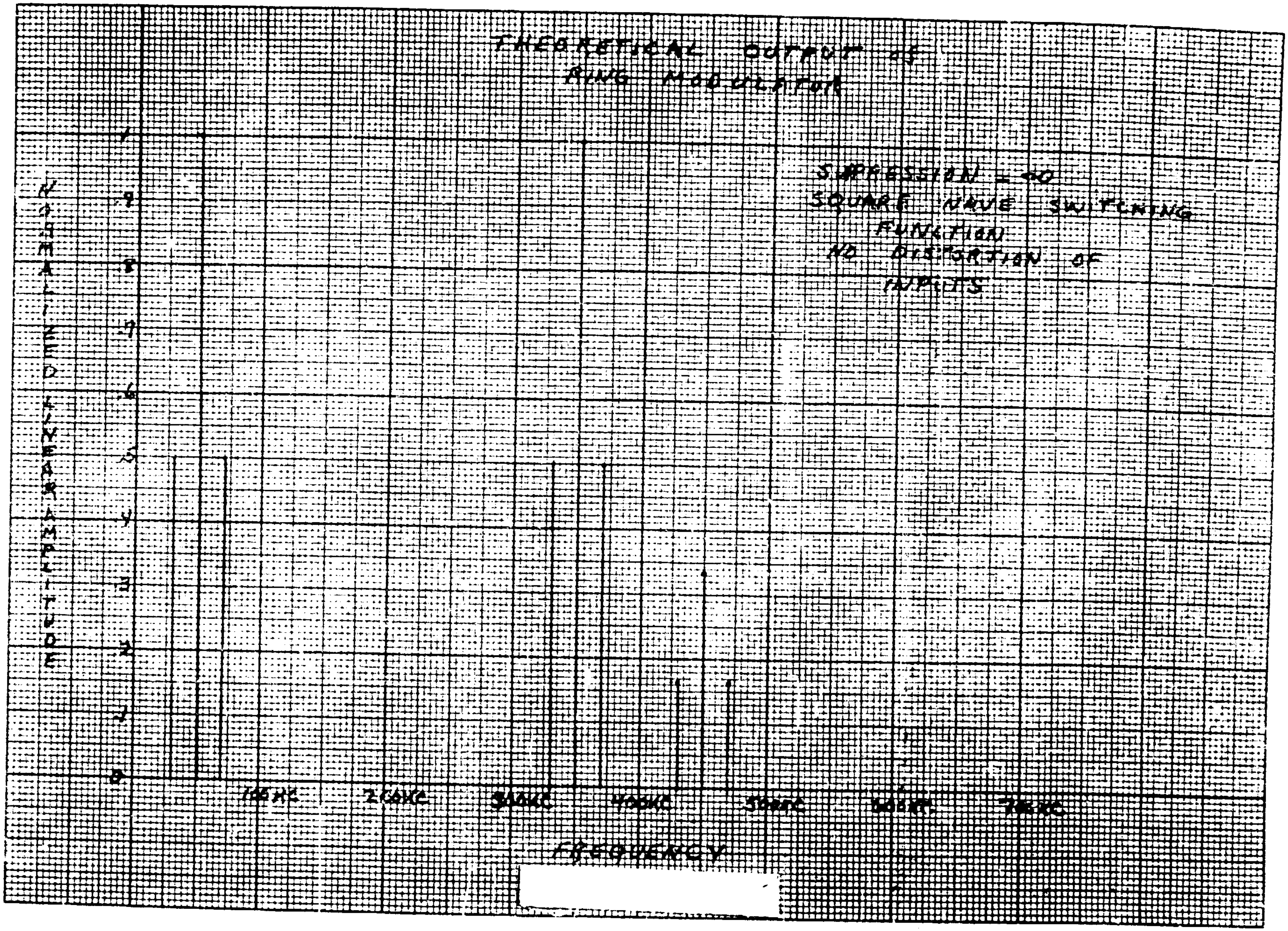


Figure 50 Theoretical Output of Ring Modulator

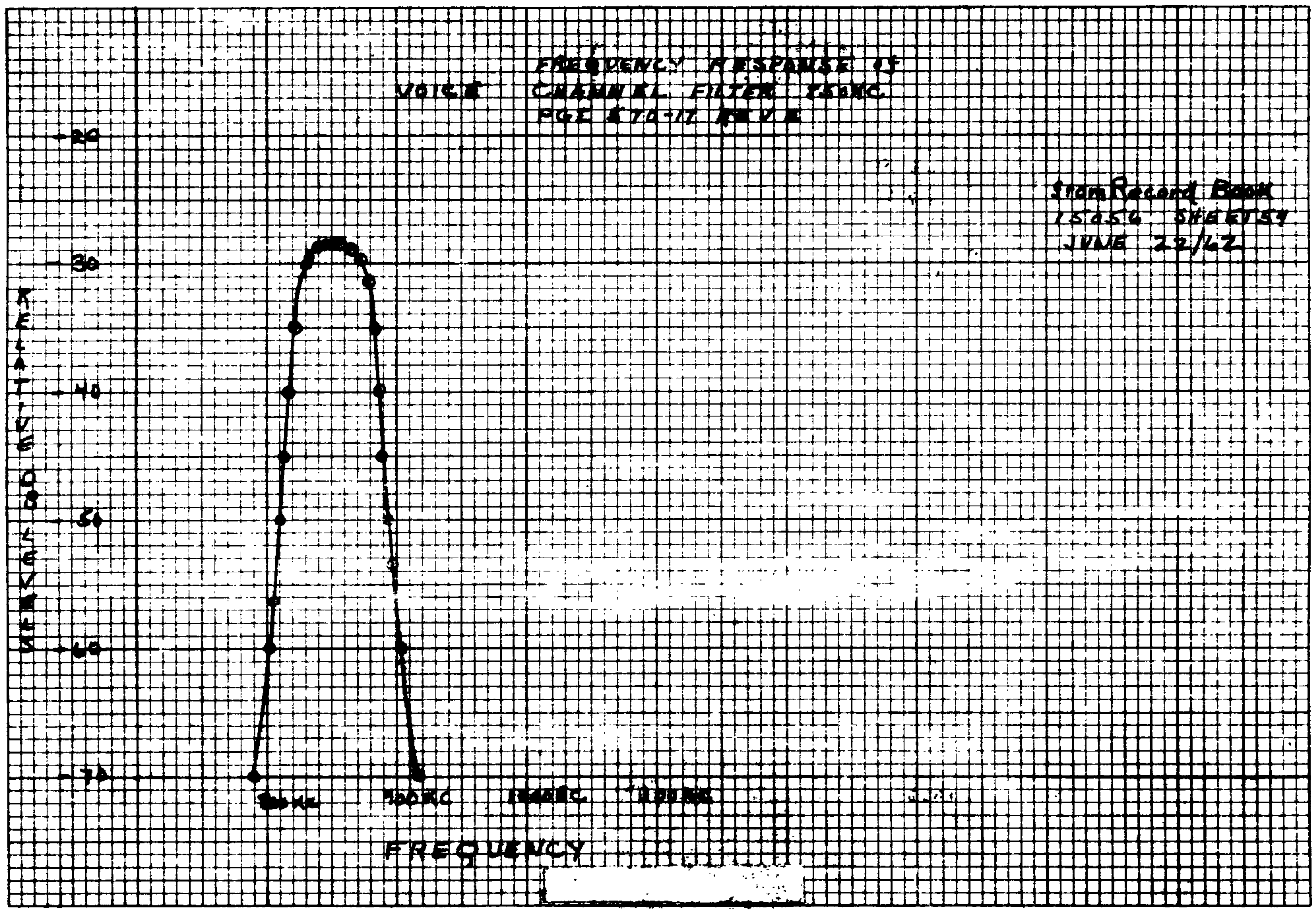


Figure 51 Frequency Response of Voice Channel Filter, 850 Kc

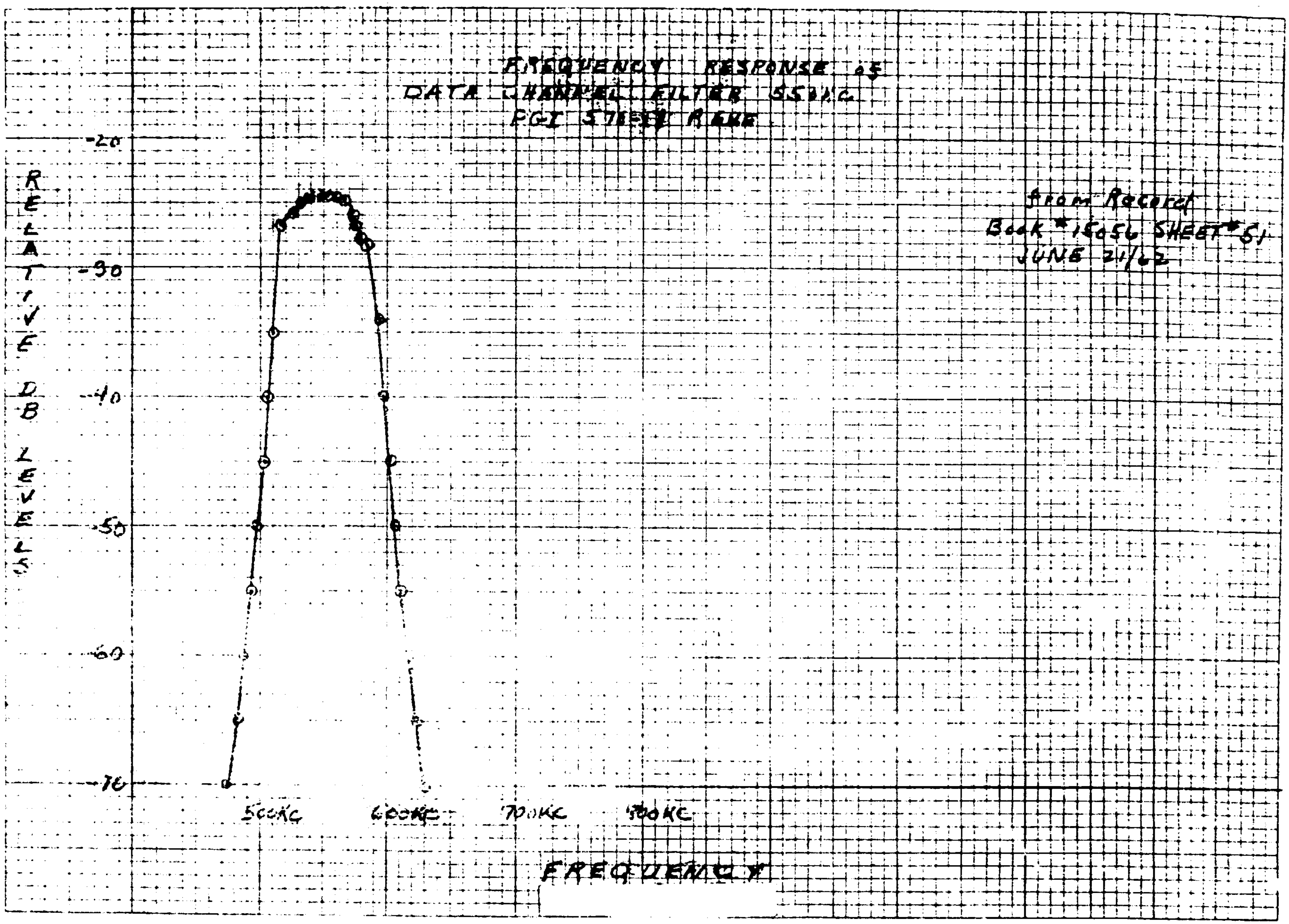


Figure 52 Frequency Response of Data Channel Filter, 550 Kc

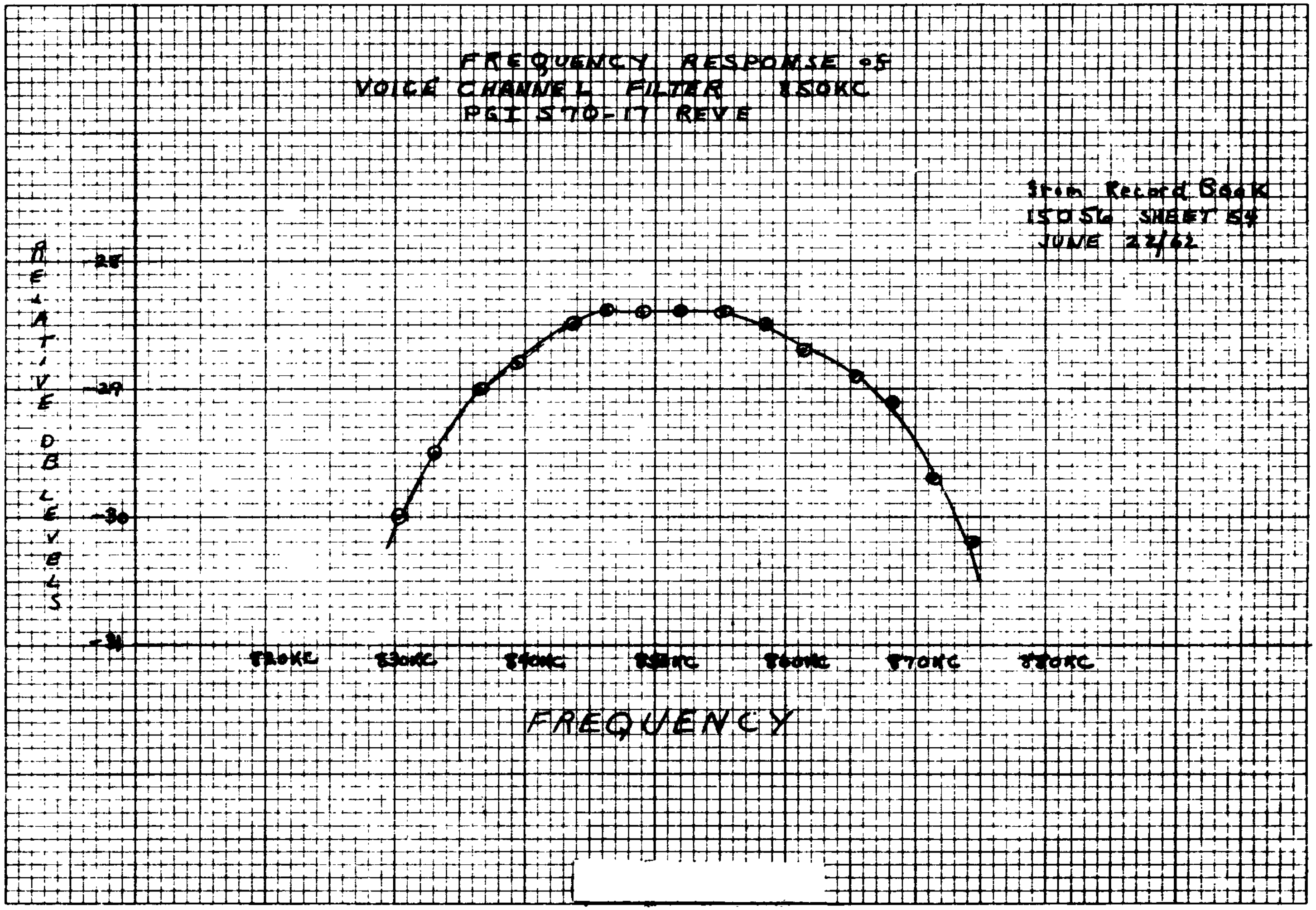


Figure 53 Frequency Response of Voice Channel Filter, 850 Kc

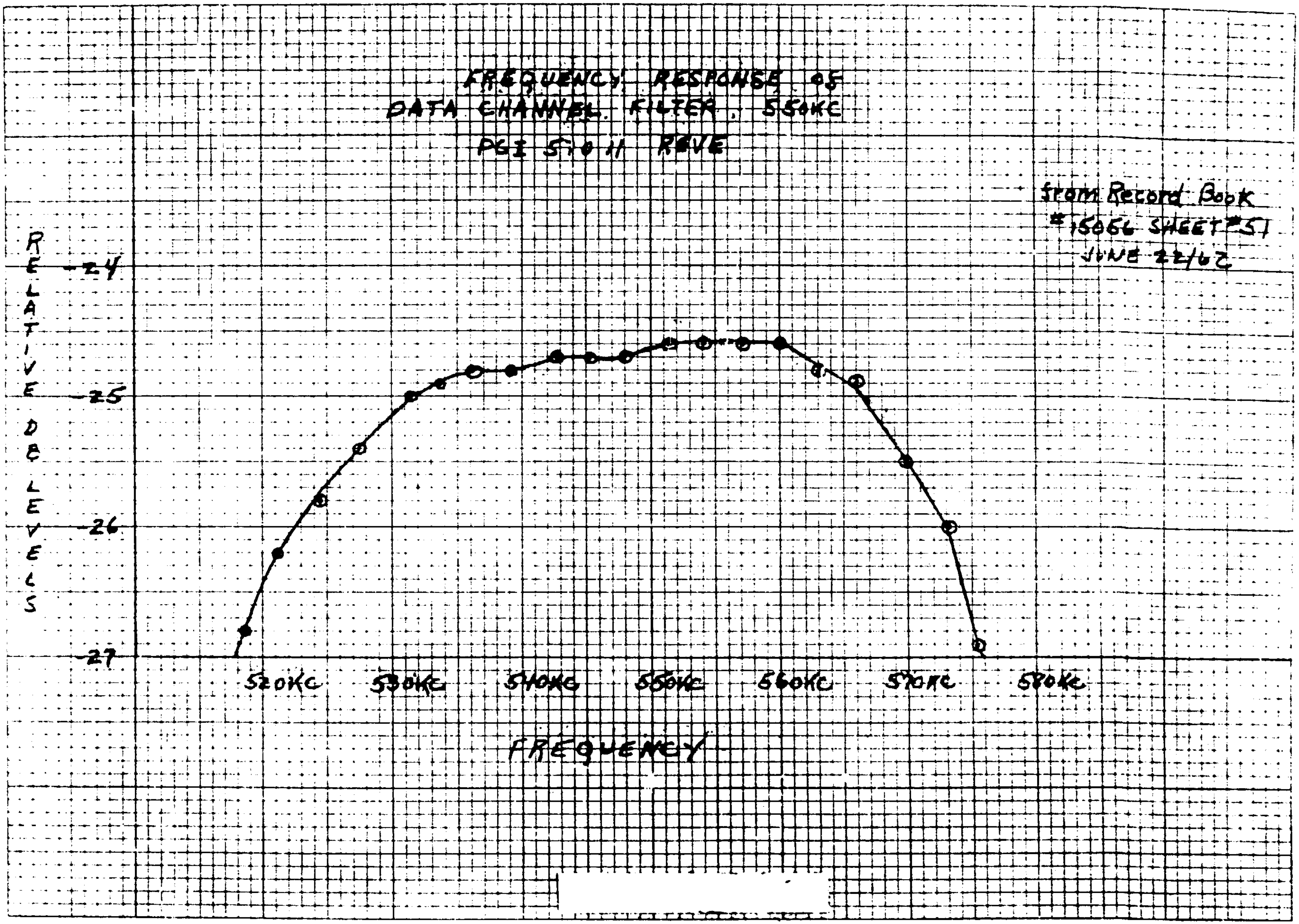


Figure 54 Frequency Response of Data Channel Filter, 550 Kc

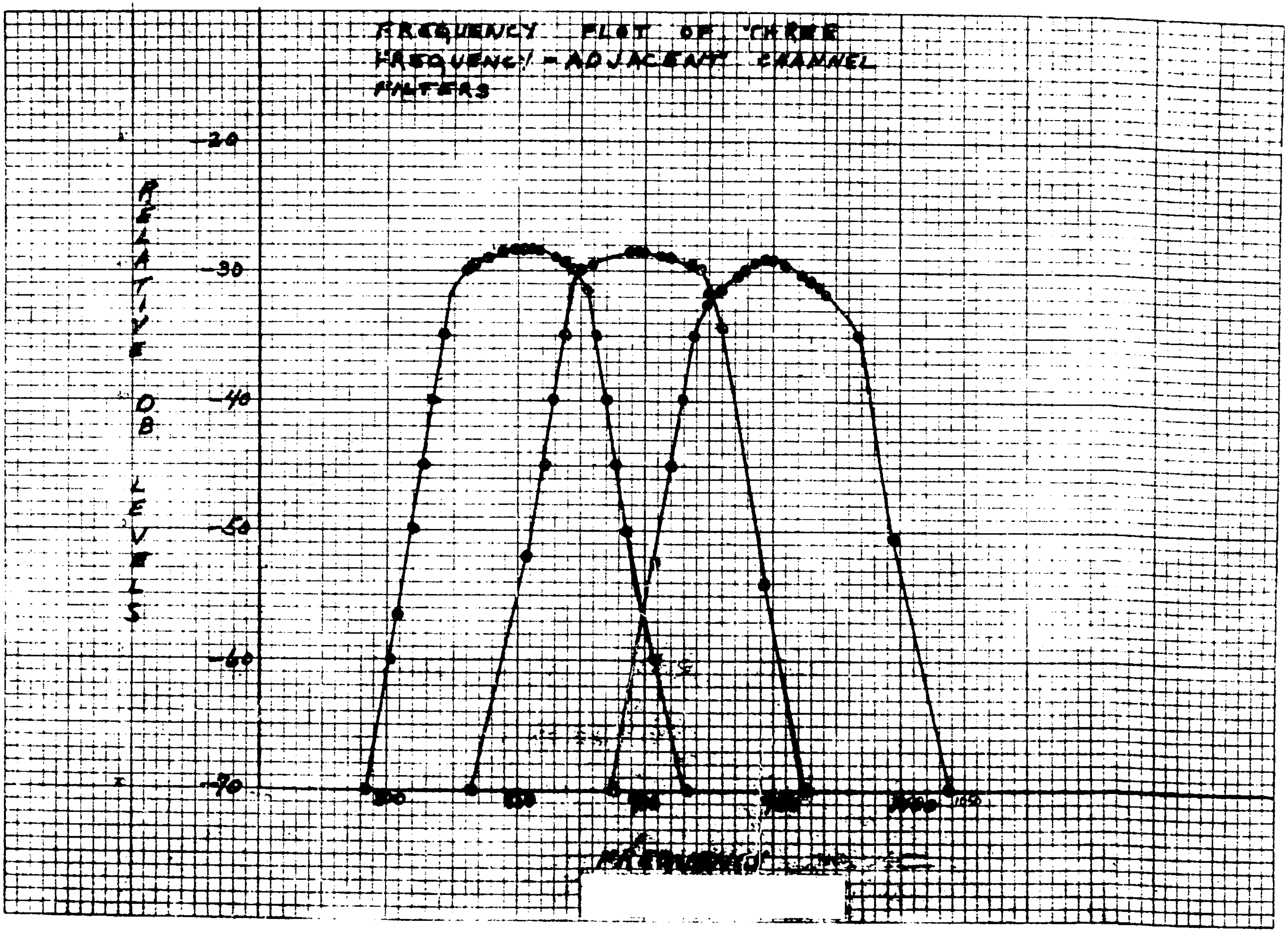


Figure 55 Frequency Plot of Three Frequency-Adjacent Channel Filters

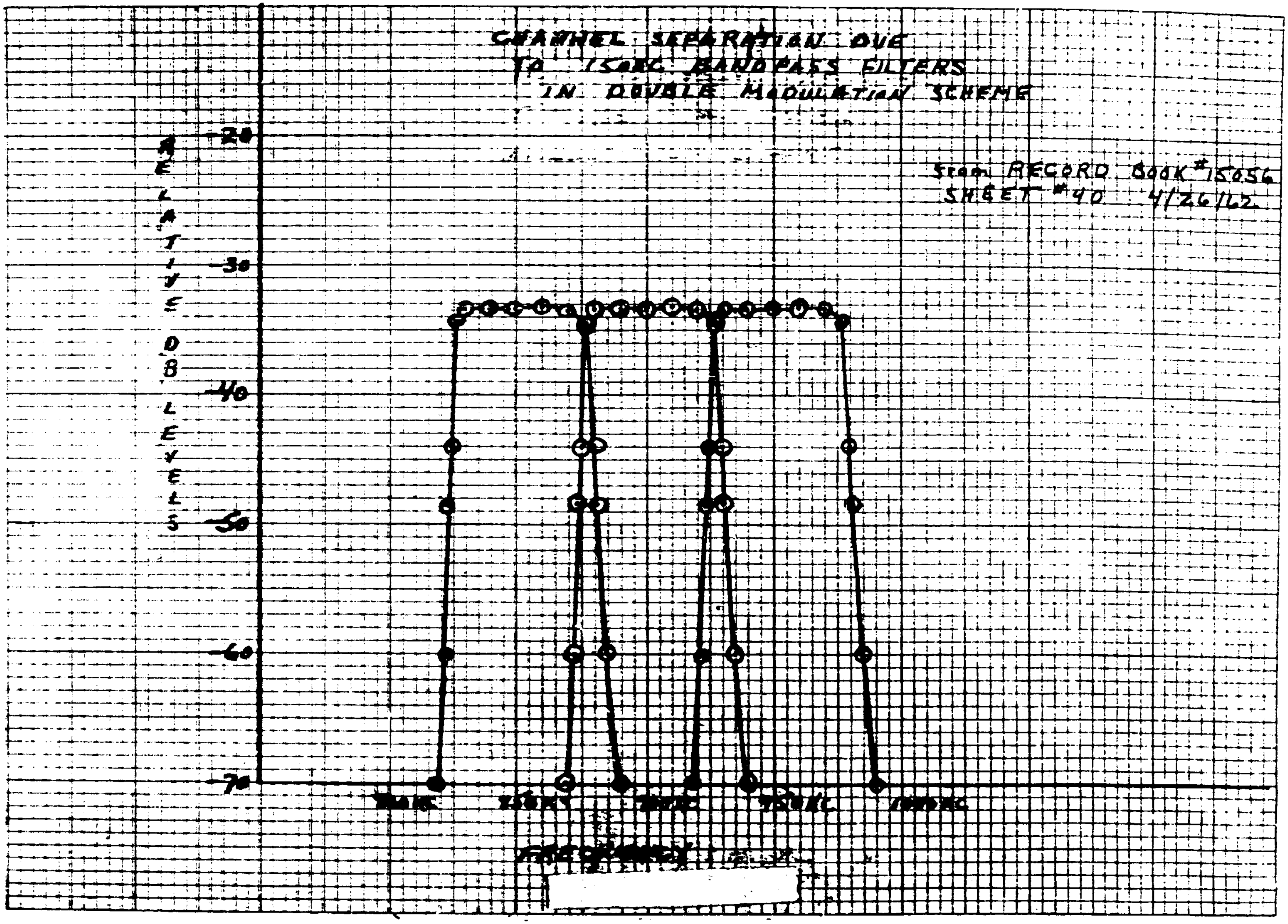


Figure [illegible]. Channel Separation Due to 150-Kc Bandpass Filters in Double-Modulation Scheme

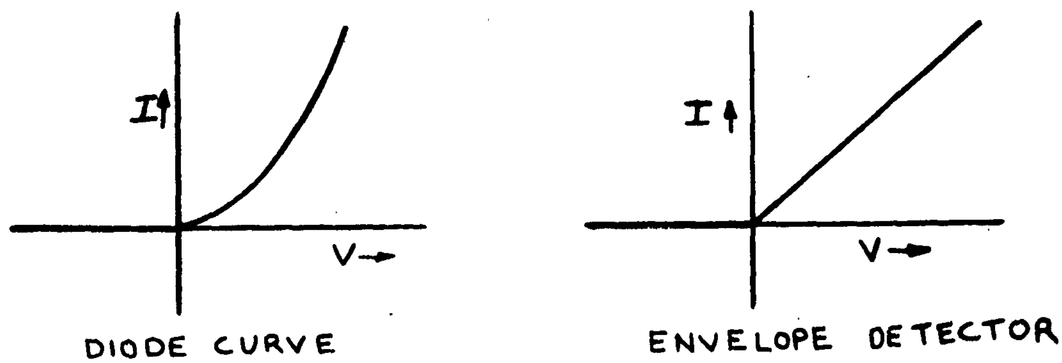


Figure 57 Diode Curve and Envelope Detector Curve

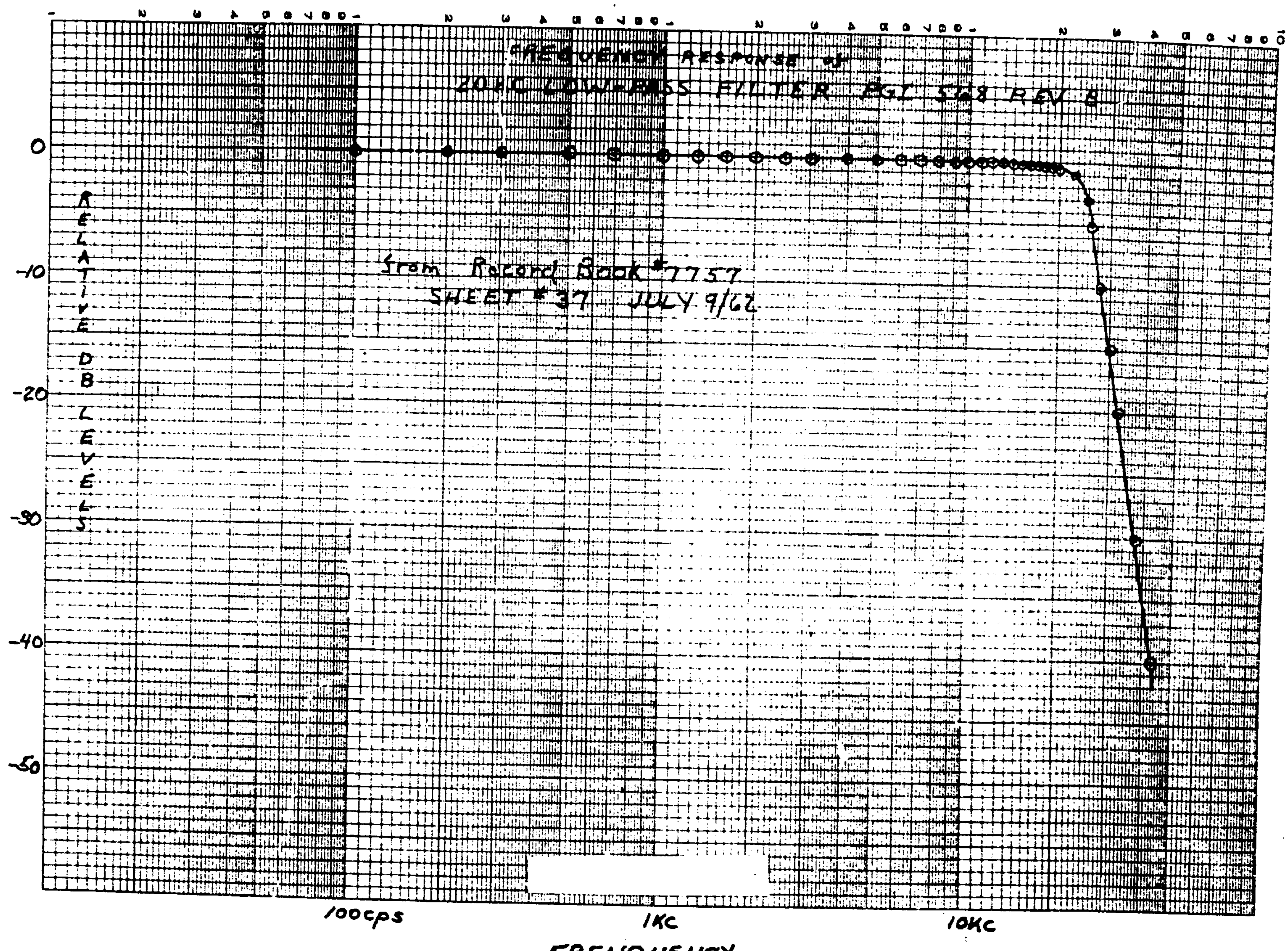


Figure 58 Frequency Response of 20-Kc Low-Pass Filter

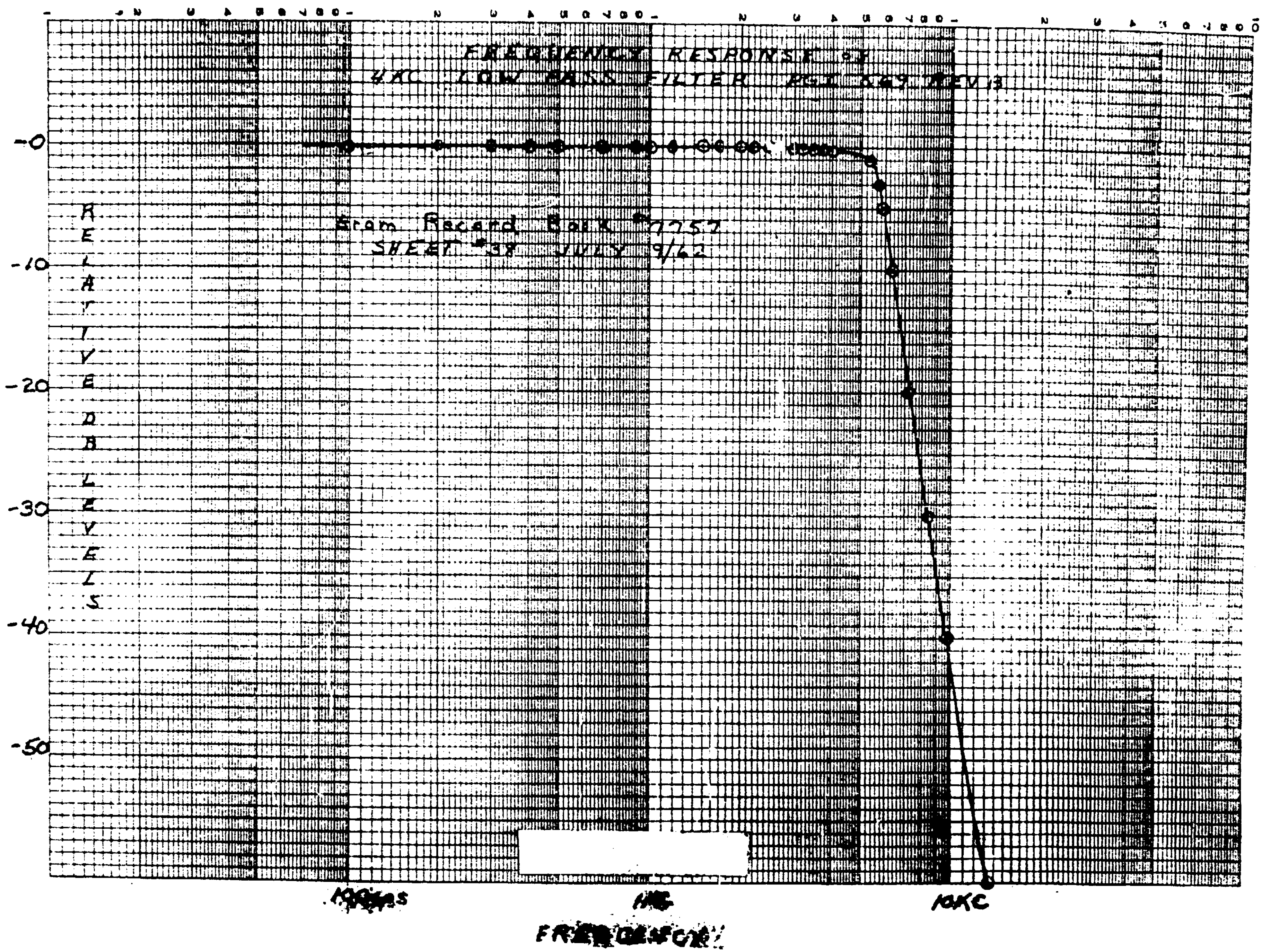


Figure 59 Frequency Response of 4-Kc Low-Pass Filter

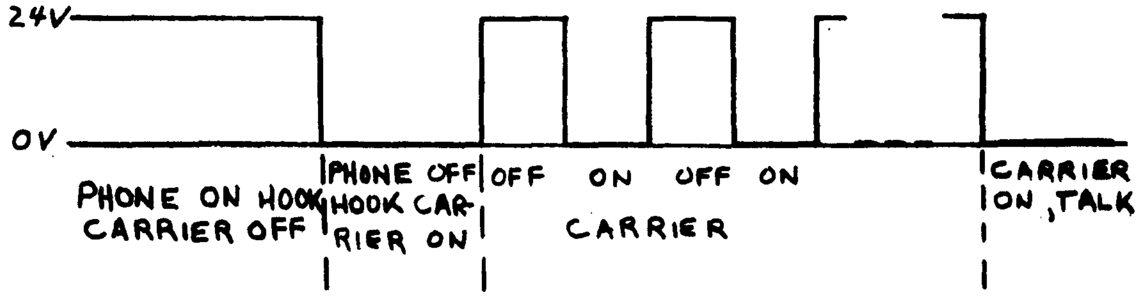
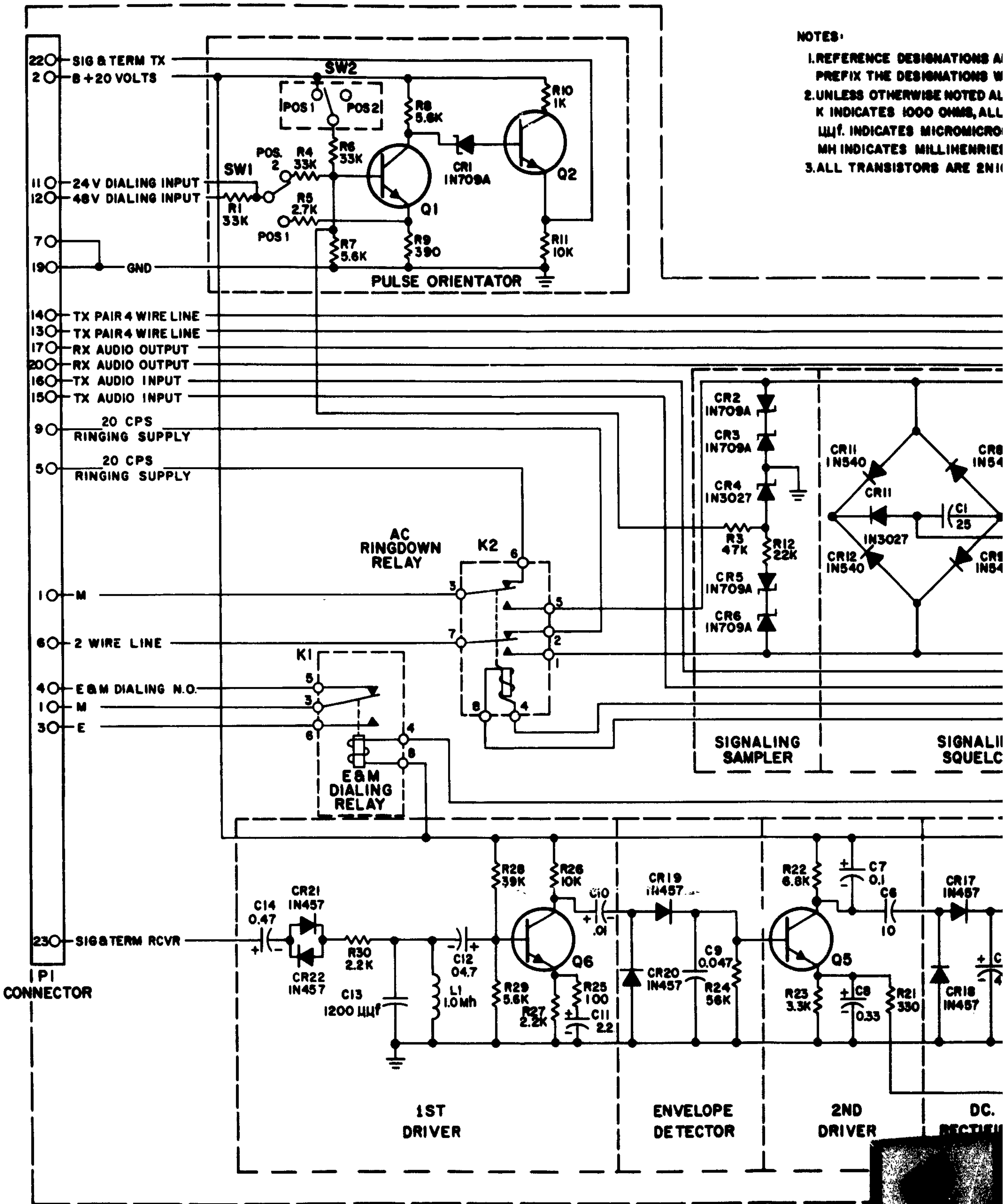


Figure 60 Dialing Sequence

NOTES:

1. REFERENCE DESIGNATIONS A
   PREFIX THE DESIGNATIONS W
2. UNLESS OTHERWISE NOTED AL
   K INDICATES 1000 OHMS, ALL
   µµf. INDICATES MICROMICRO
   MH INDICATES MILLIHENRIE
3. ALL TRANSISTORS ARE 2N1



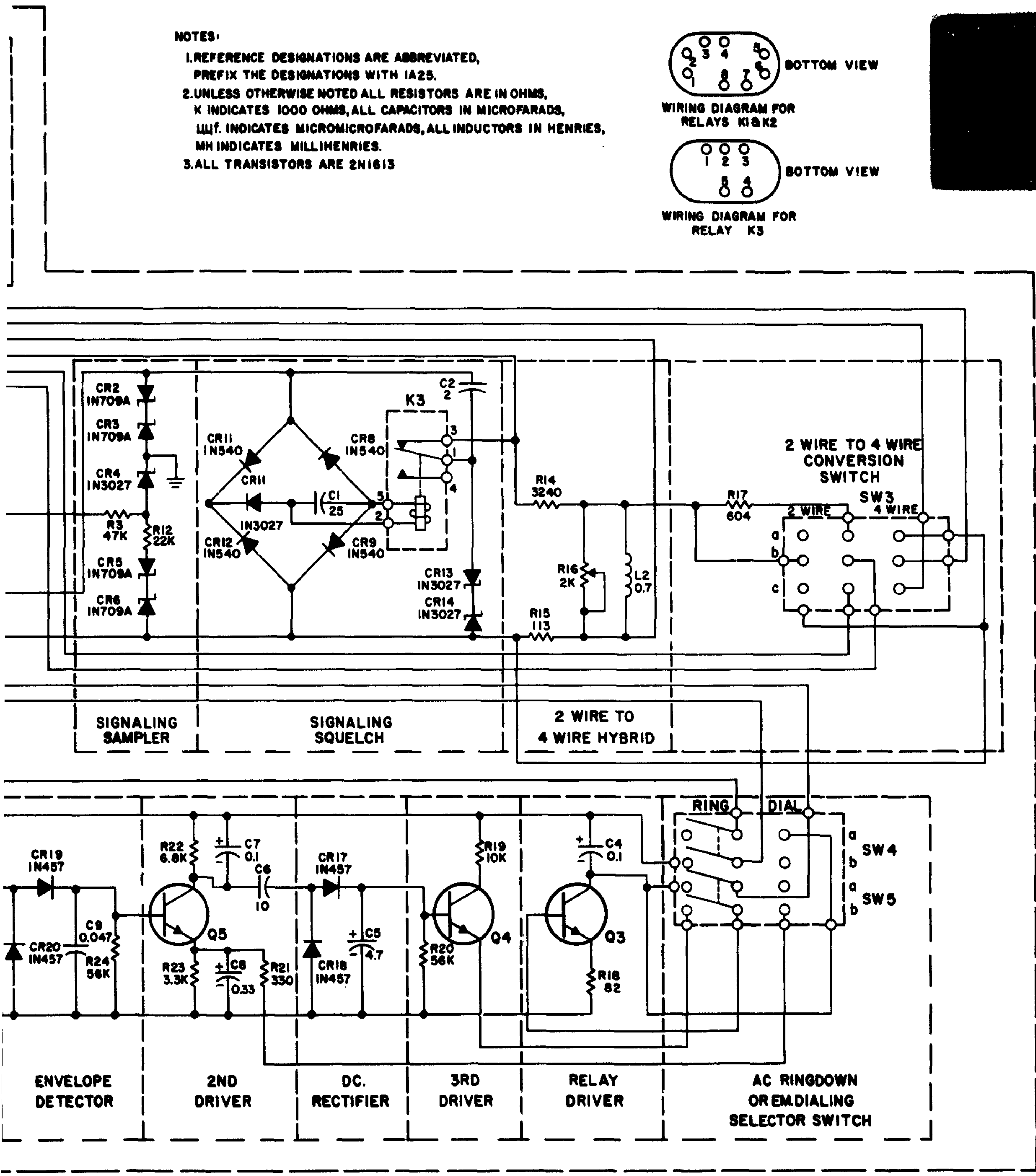


Figure 61 Signaling and Termination Assembly, Schematic Diagram

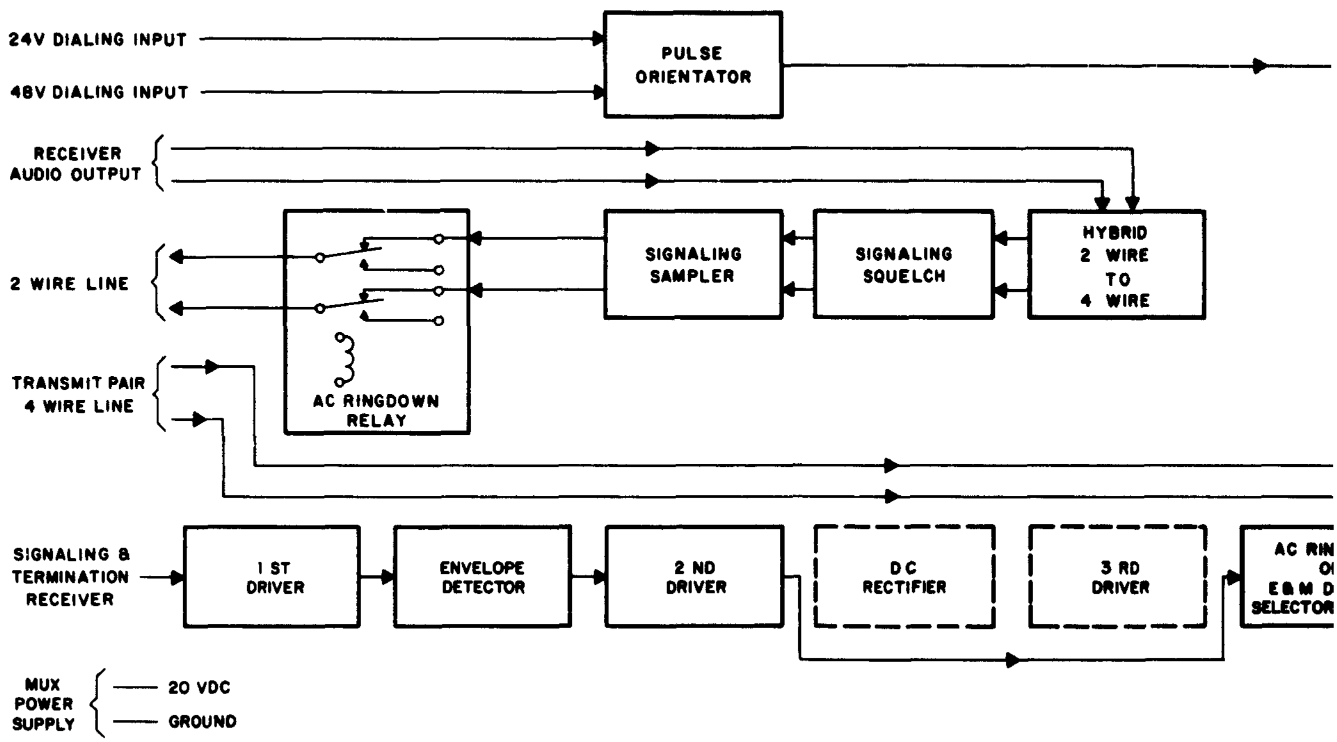


Figure 62 Signa
Bloc

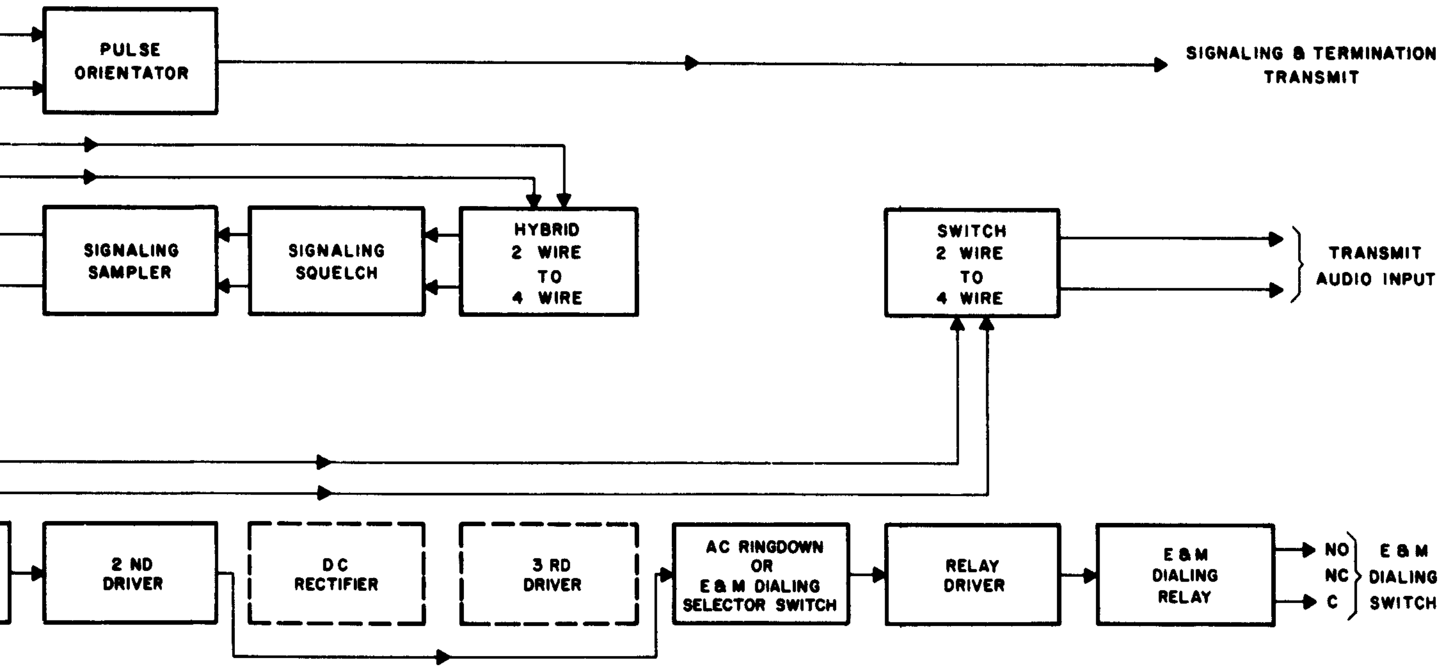


Figure 62 Signaling and Termination Assembly, E & M Dialing, Block Diagram

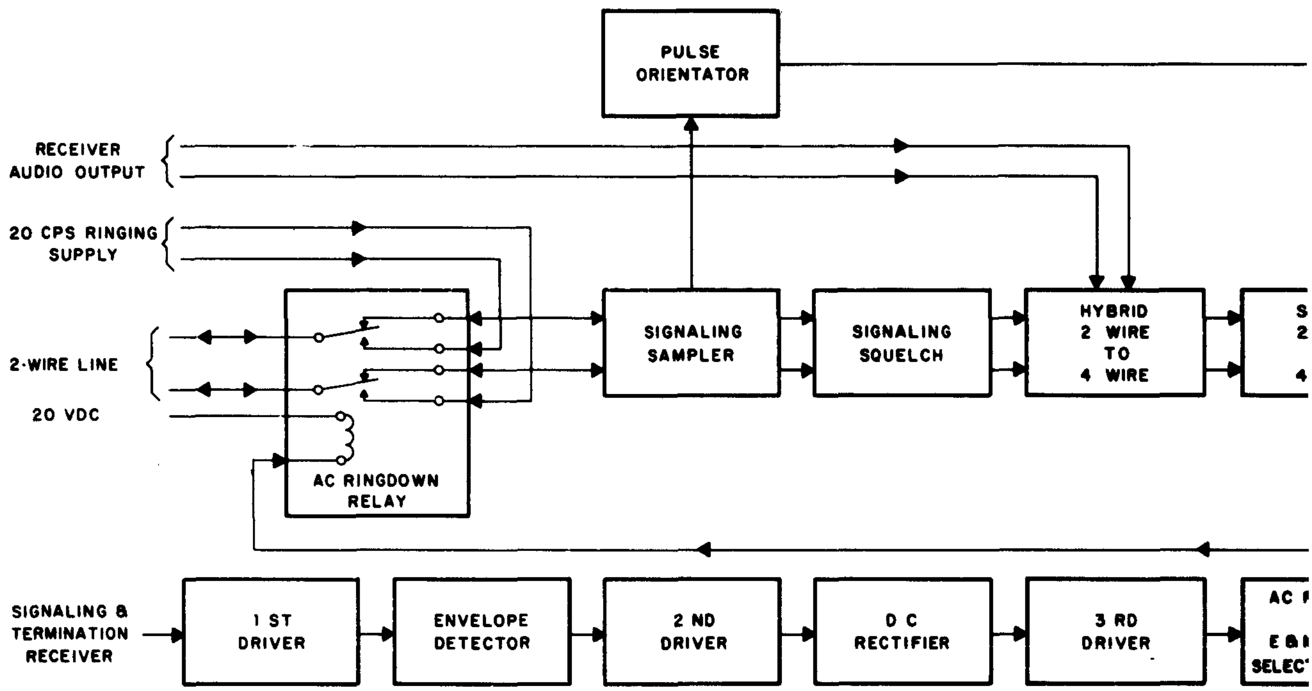


Figure 63 Sig
Bl

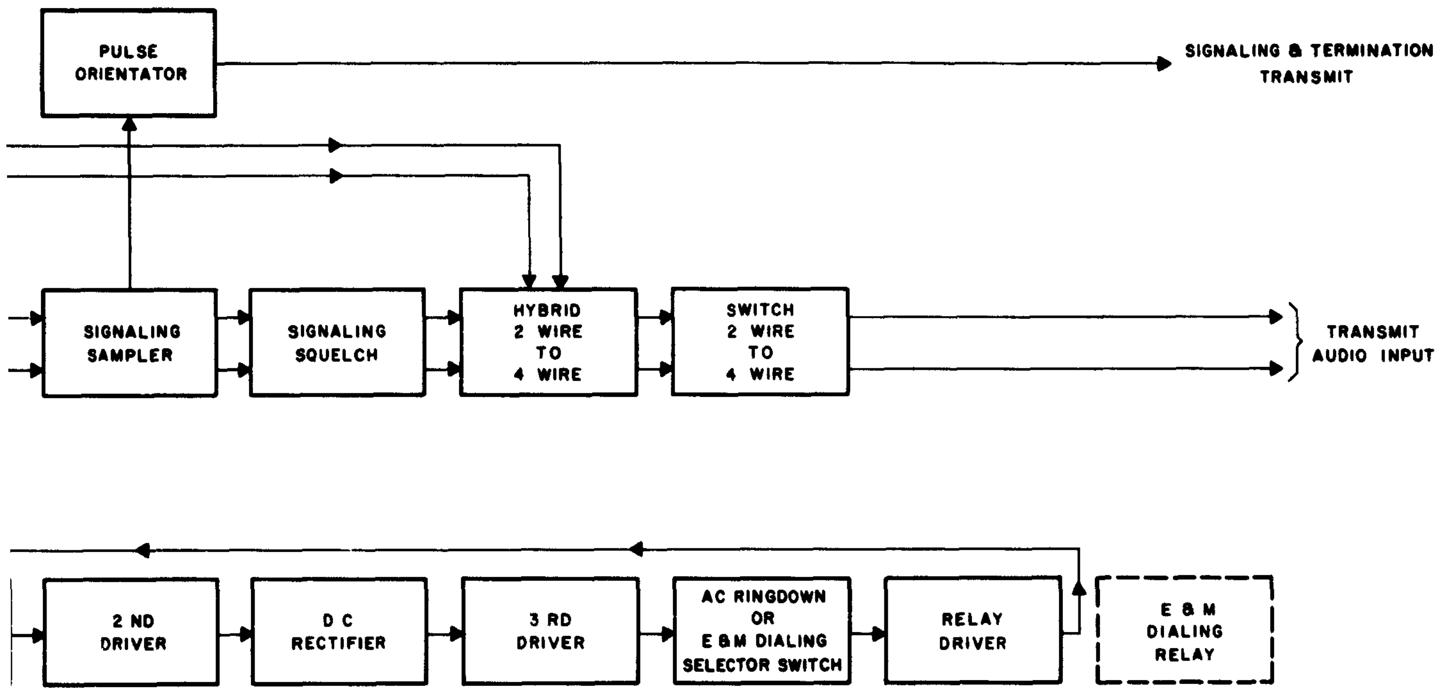


Figure 63 Signaling and Termination Assembly, AC Ringdown, Block Diagram

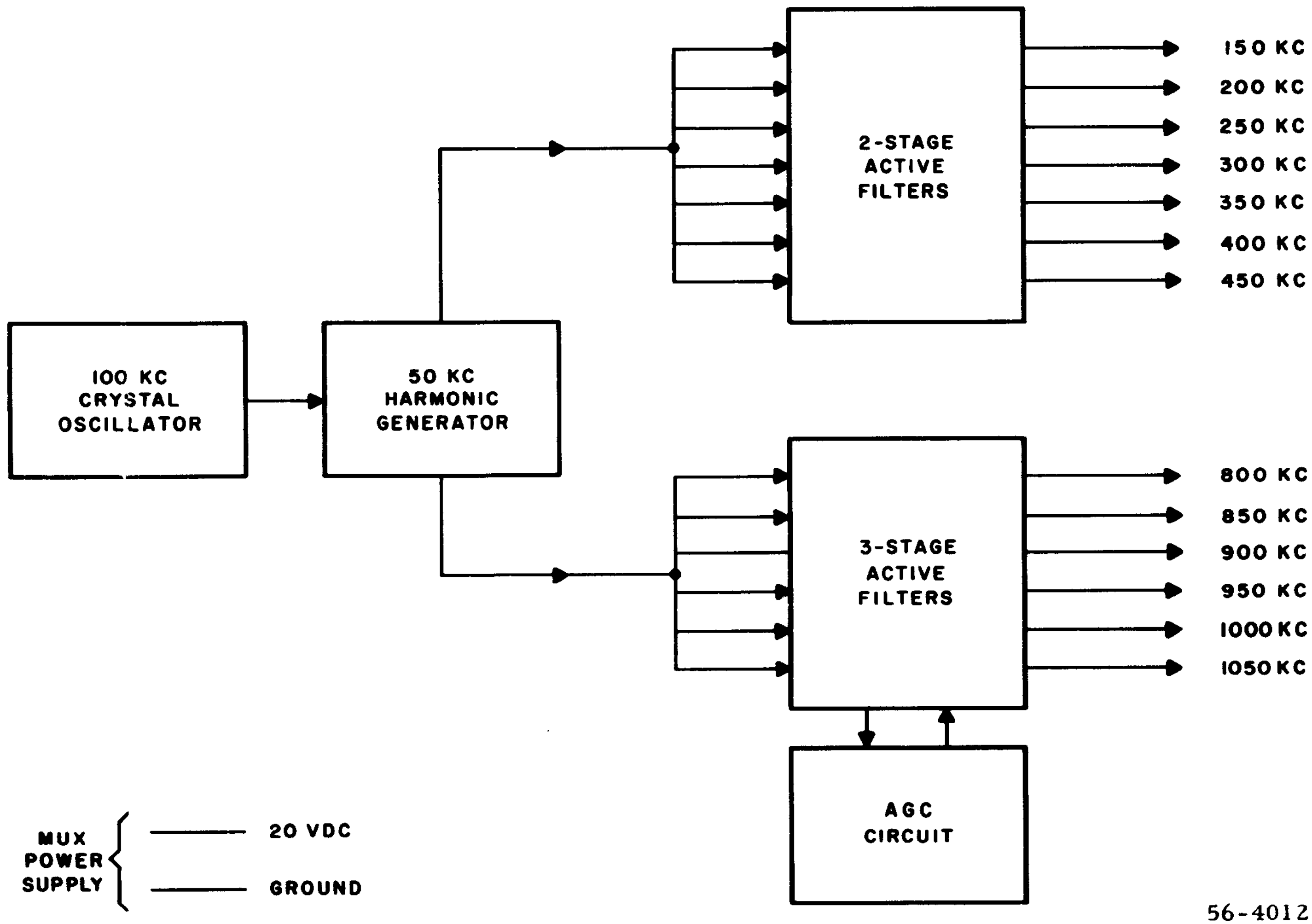
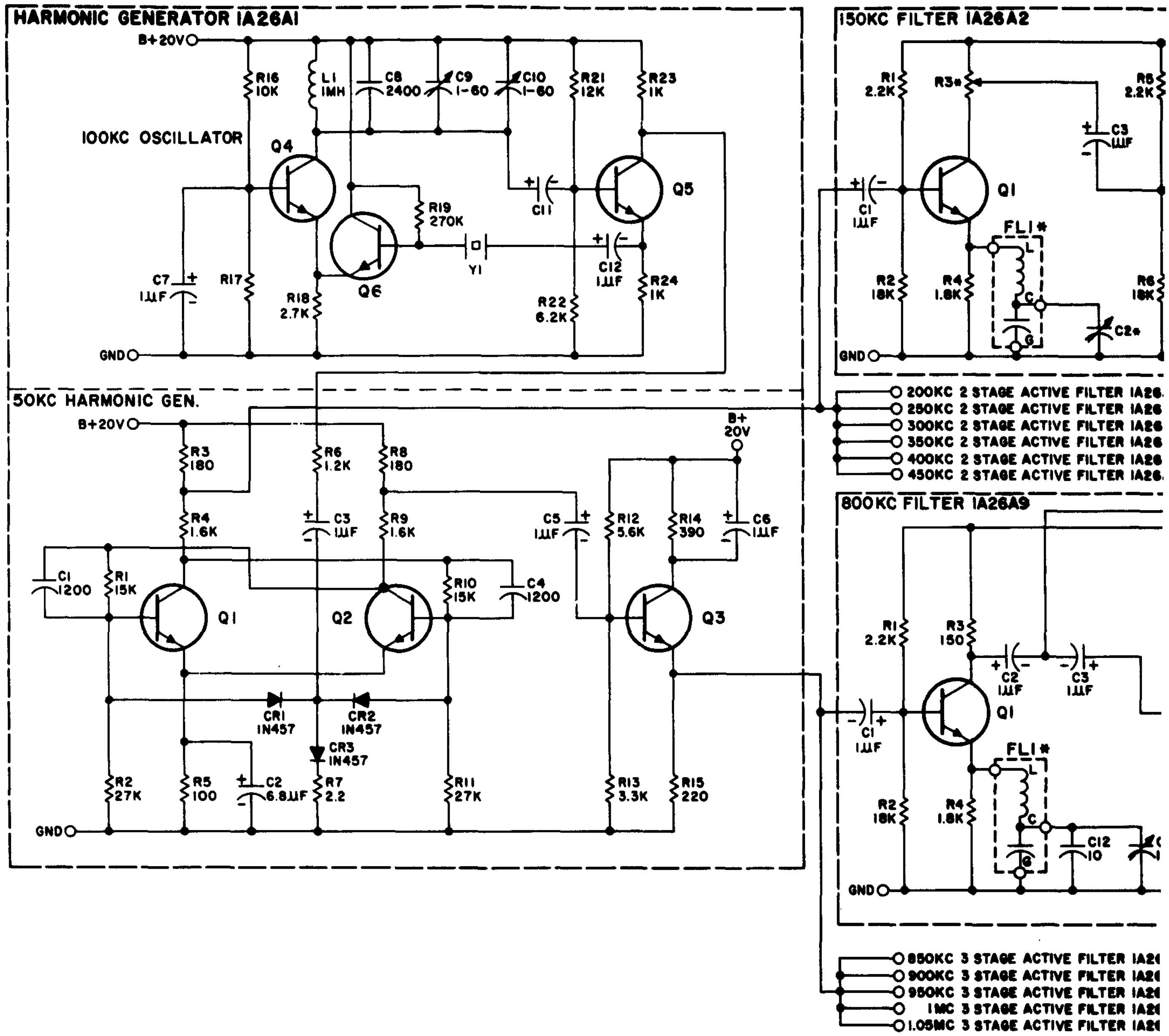
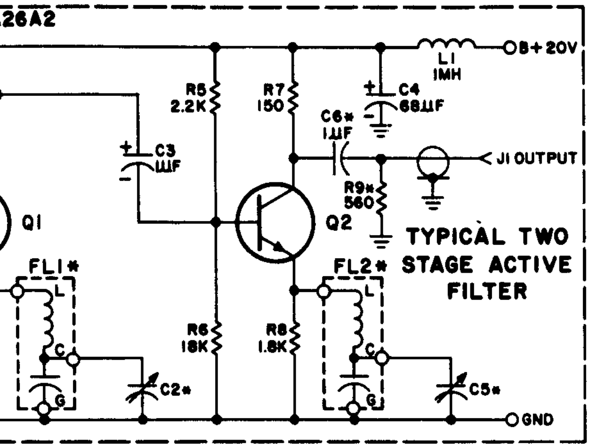


Figure 64 Carrier Generator, Block Diagram



HARMONIC GENERATOR 1A26A1
100KC OSCILLATOR
50KC HARMONIC GEN.
150KC FILTER 1A26A2
800KC FILTER 1A26A9
200KC 2 STAGE ACTIVE FILTER 1A26
250KC 2 STAGE ACTIVE FILTER 1A26
300KC 2 STAGE ACTIVE FILTER 1A26
350KC 2 STAGE ACTIVE FILTER 1A26
400KC 2 STAGE ACTIVE FILTER 1A26
450KC 2 STAGE ACTIVE FILTER 1A26
850KC 3 STAGE ACTIVE FILTER 1A2
900KC 3 STAGE ACTIVE FILTER 1A2
950KC 3 STAGE ACTIVE FILTER 1A2
1MC 3 STAGE ACTIVE FILTER 1A2
1.05MC 3 STAGE ACTIVE FILTER 1A2

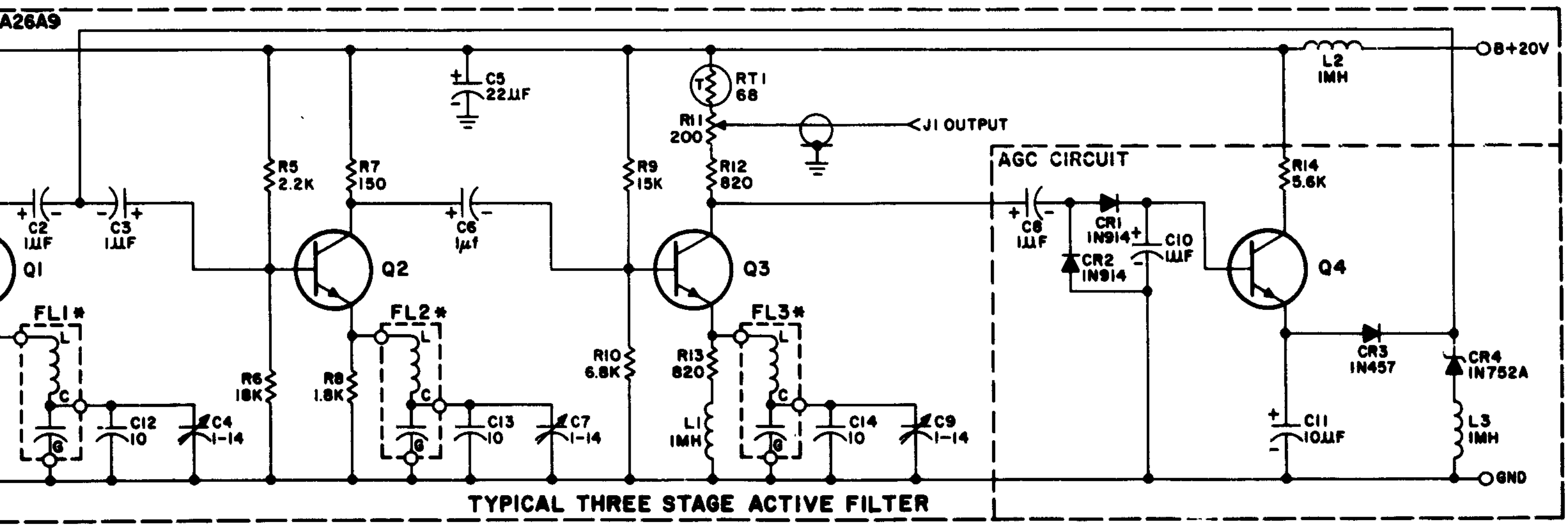
1



STAGE ACTIVE FILTER IA26A3
STAGE ACTIVE FILTER IA26A4
STAGE ACTIVE FILTER IA26A5
STAGE ACTIVE FILTER IA26A6
STAGE ACTIVE FILTER IA26A7
STAGE ACTIVE FILTER IA26A8

REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE APPROPRIATE REFERENCE DESIGNATION FROM THE SUBASSEMBLY.

ALL RESISTOR VALUES ARE OHMS UNLESS OTHERWISE MARKED.
K=1000 OHMS
ALL TRANSISTORS ARE 2N1613'S.
ALL CAPACITORS ARE MICROMICROFARADS UNLESS OTHERWISE MARKED.
µF INDICATES MICROFARADS.
MH INDICATES MILLIHENRIES.
*COMPONENTS VARY WITH FREQUENCY SEE PARTS LIST.
C-12,13,14 ARE USED WHERE NEEDED TO EXTEND TUNING RANGE OF C-4,7 AND/OR 9.
C6 AND R9 ARE USED ONLY ON THE 150KC TWO STAGE ACTIVE FILTER CARD.



STAGE ACTIVE FILTER IA26A10
STAGE ACTIVE FILTER IA26A11
STAGE ACTIVE FILTER IA26A12
STAGE ACTIVE FILTER IA26A13
STAGE ACTIVE FILTER IA26A14

Figure 65 Carrier Generator, Schematic Diagram

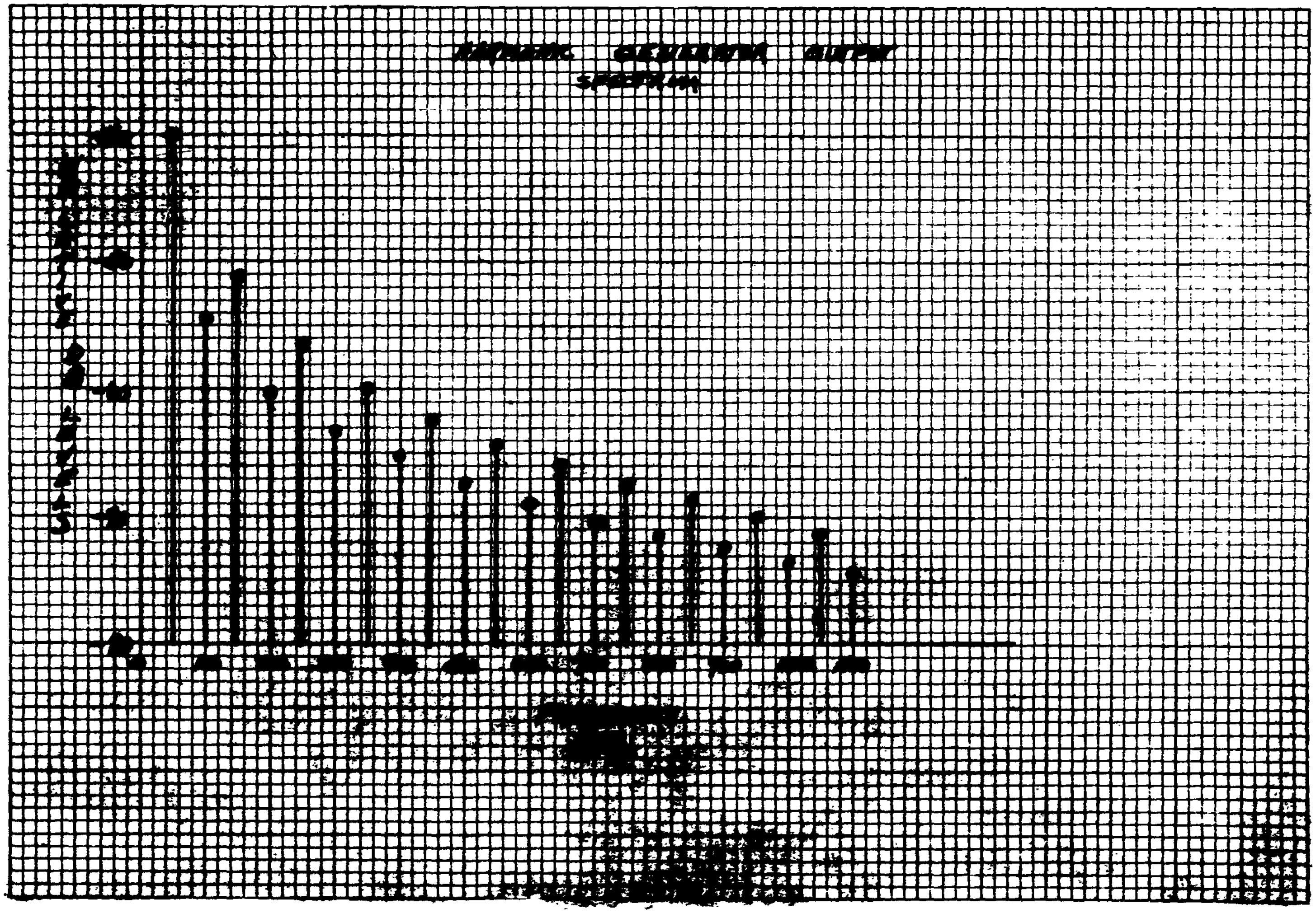


Figure 66 Harmonic Generator Output Spectrum

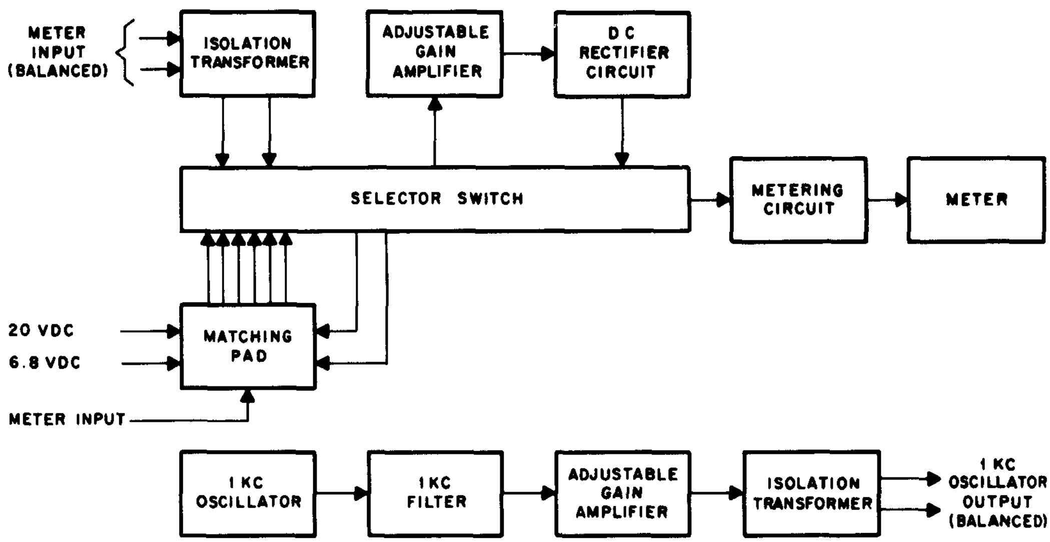
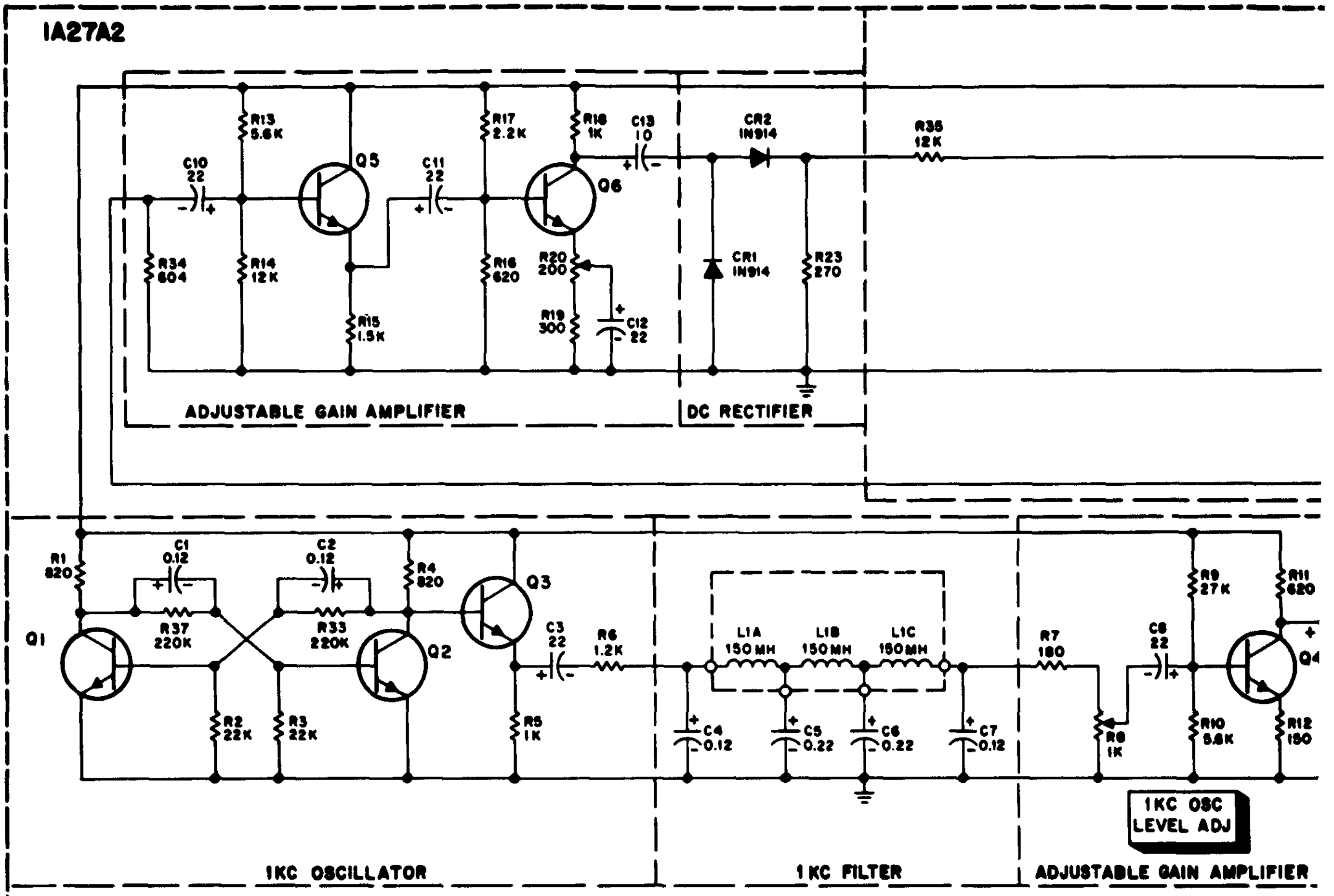


Figure 67 Metering and Test Facility, Block Diagram

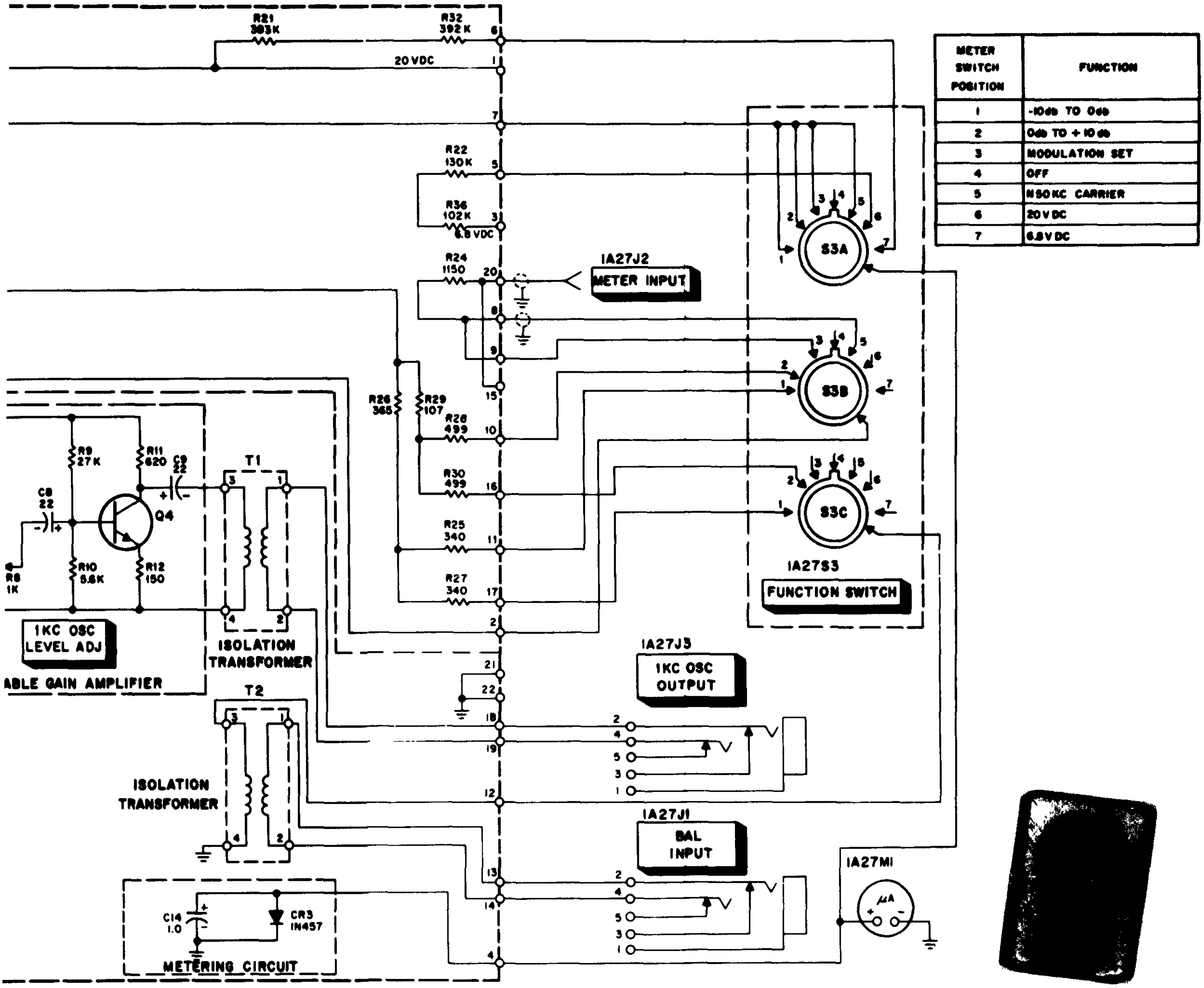


NOTE 1.
REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE DESIGNATIONS WITH 1A27A2 EXCEPT WHERE NOTED WITH AN ASTERISK.

NOTE 2.
UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS ALL CAPACITANCE VALUES ARE IN MICROFARADS
K INDICATES THOUSANDS OF OHMS
MEG INDICATES MILLIONS OF OHMS
μμF INDICATES MICROMICROFARADS
MH INDICATES MILLI HENRIES
ALL TRANSISTORS ARE 2N1613

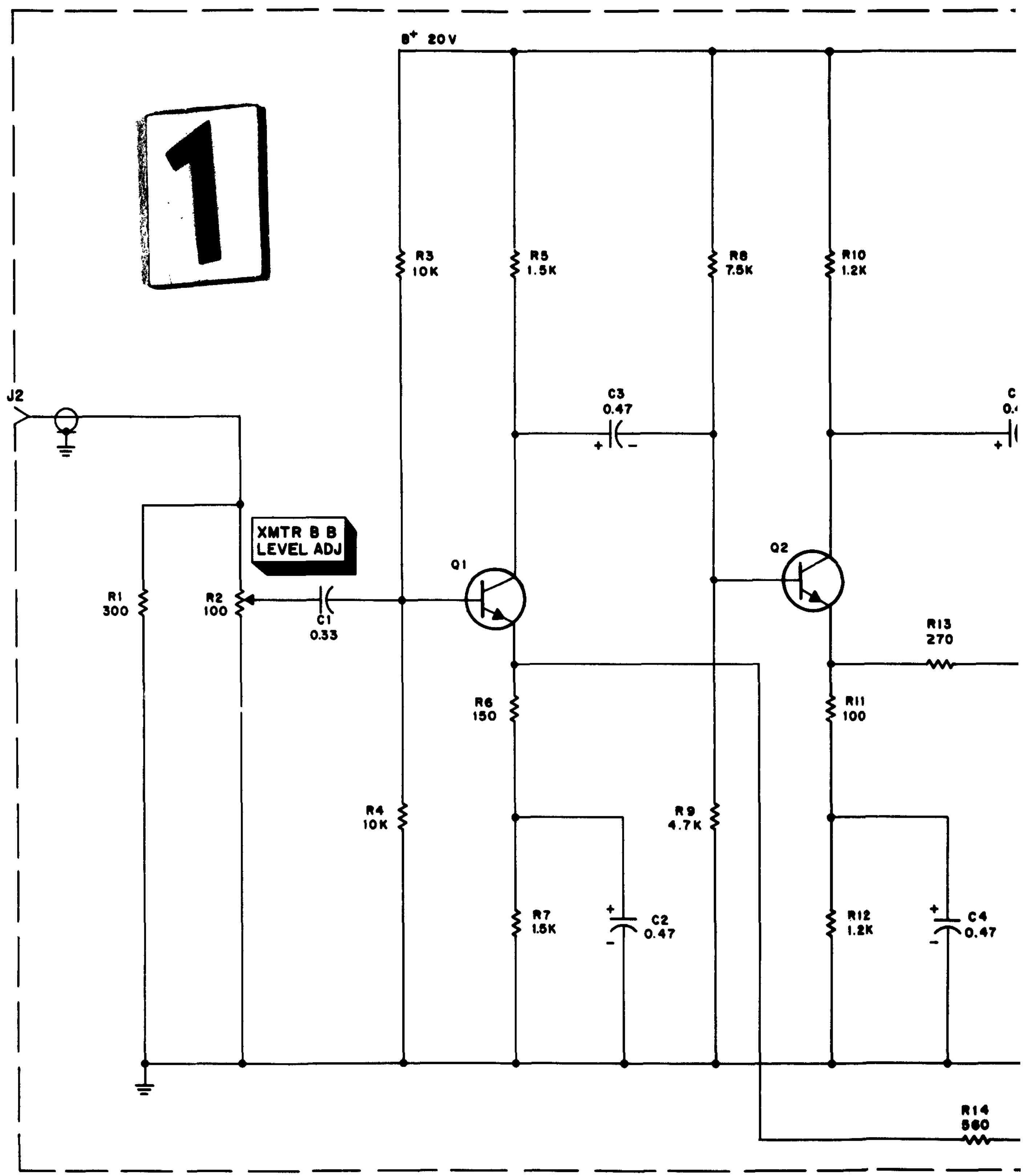
ISOL.
TRANSF

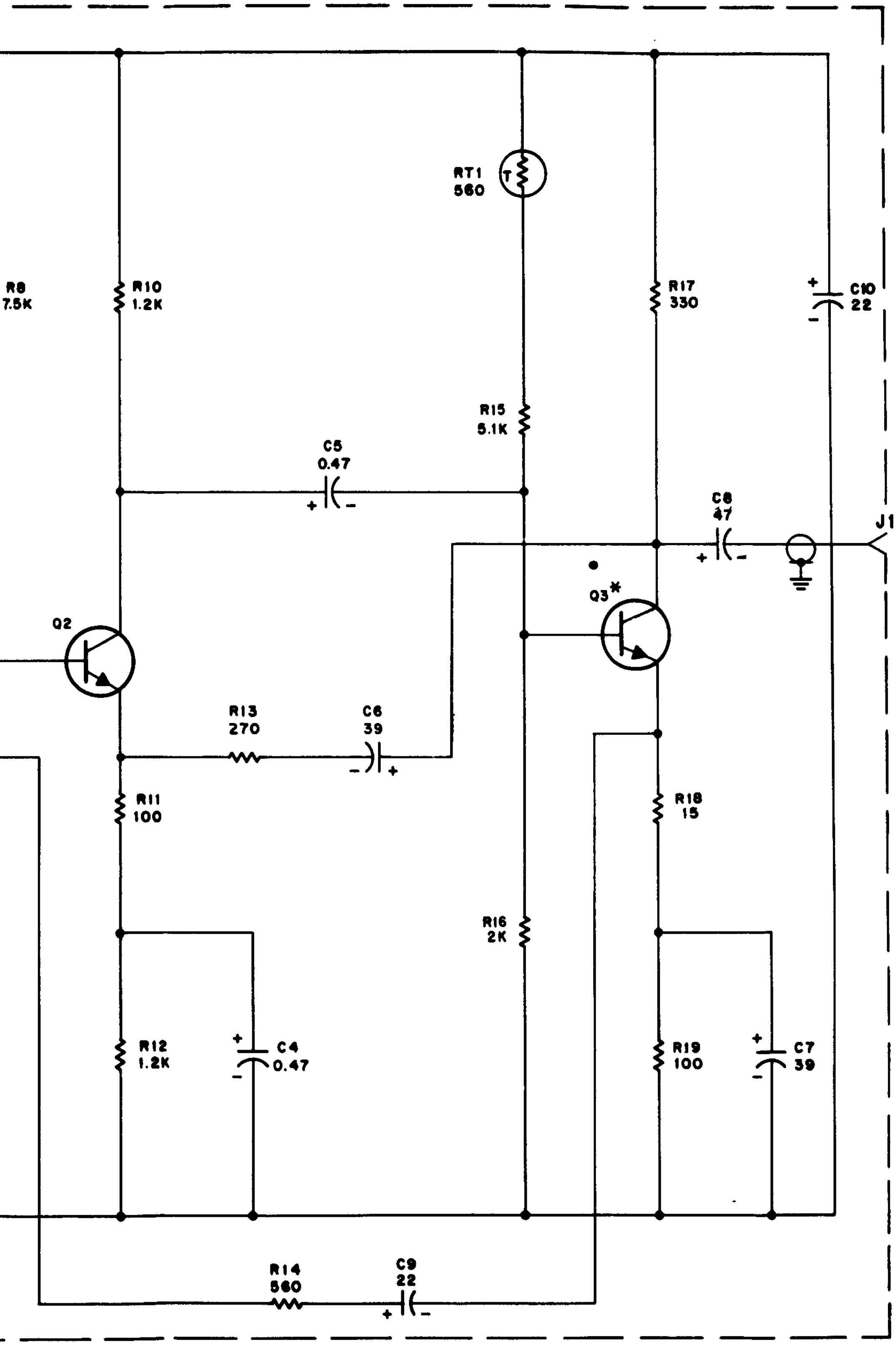
1



METER SWITCH POSITION	FUNCTION
1	-10db TO 0db
2	0db TO + 10 db
3	MODULATION SET
4	OFF
5	H 50KC CARRIER
6	20 V DC
7	6.8 V DC

Figure 68 Metering and Test Facility, Schematic Diagram


1
B+ 20V
J2
R1
300
R2
100
XMTR B B
LEVEL ADJ
C1
0.33
R3
10K
R4
10K
Q1
R5
1.5K
R6
150
R7
1.5K
C2
0.47
C3
0.47
R8
7.5K
R9
4.7K
Q2
R10
1.2K
R11
100
R12
1.2K
C4
0.47
R13
270
R14
560



NOTES

* HEAT SINK IS USED

1. Q1, Q2, Q3 ARE 2N1613
UNLESS OTHERWISE STATED,
ALL RESISTORS ARE IN OHMS.
ALL CAPACITORS ARE IN MICRO-
FARADS.

2. REFERENCE DESIGNATIONS
ARE ABBREVIATED. PREFIX
THE DESIGNATIONS WITH 1A27A1

Figure 69 Baseband Amplifier, Schematic Diagram

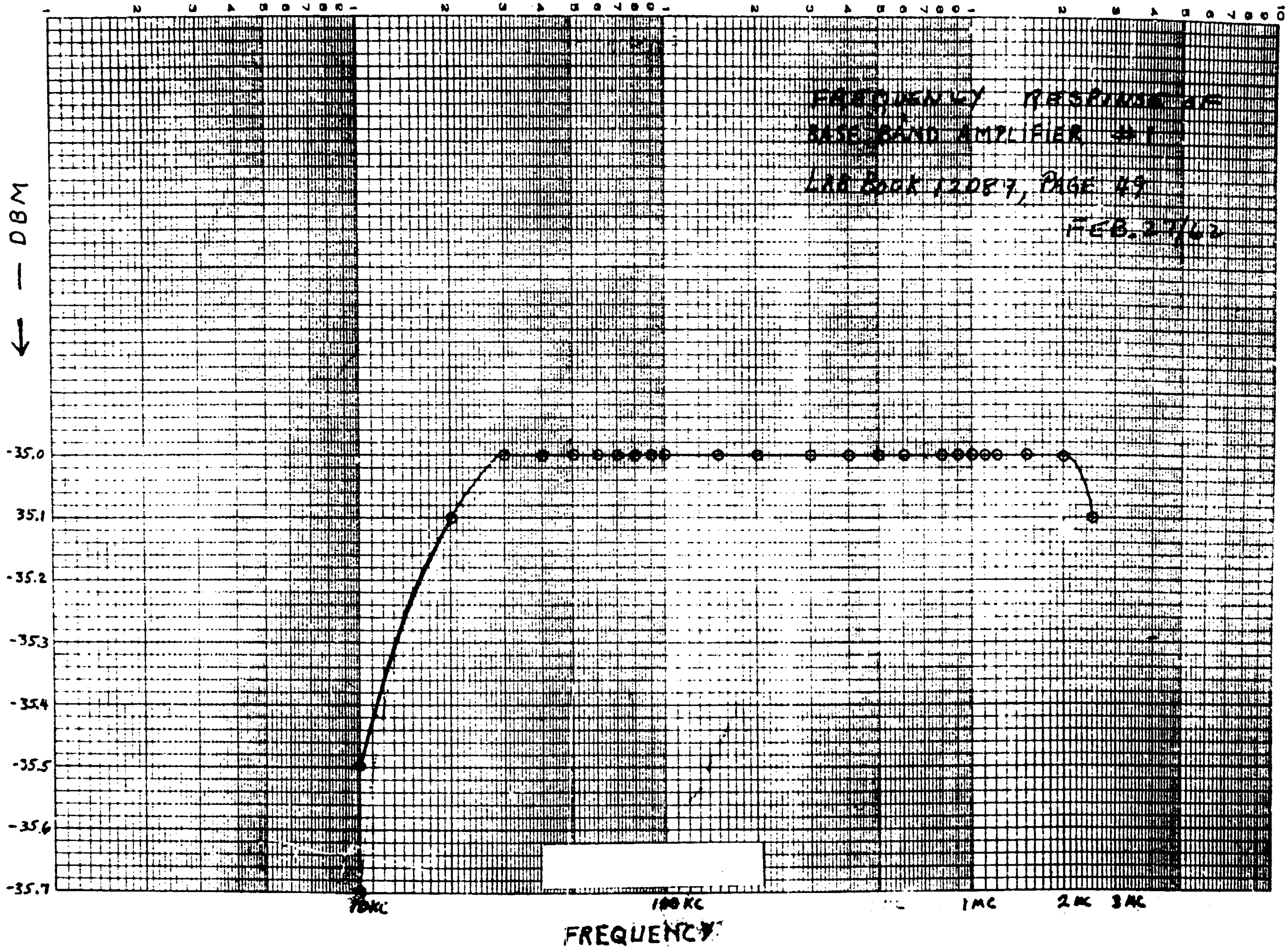


Figure 70 Frequency Response of Baseband Amplifier

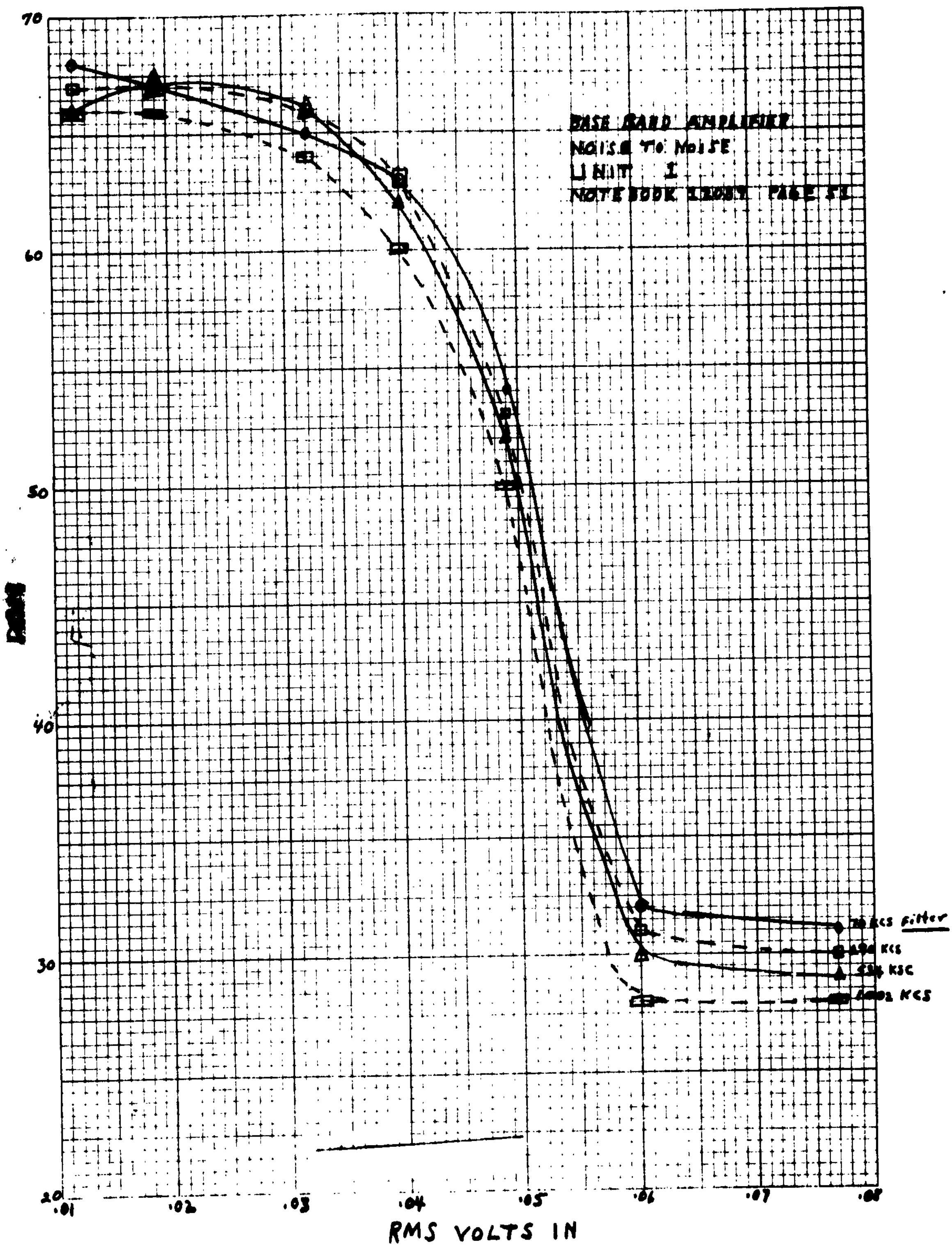
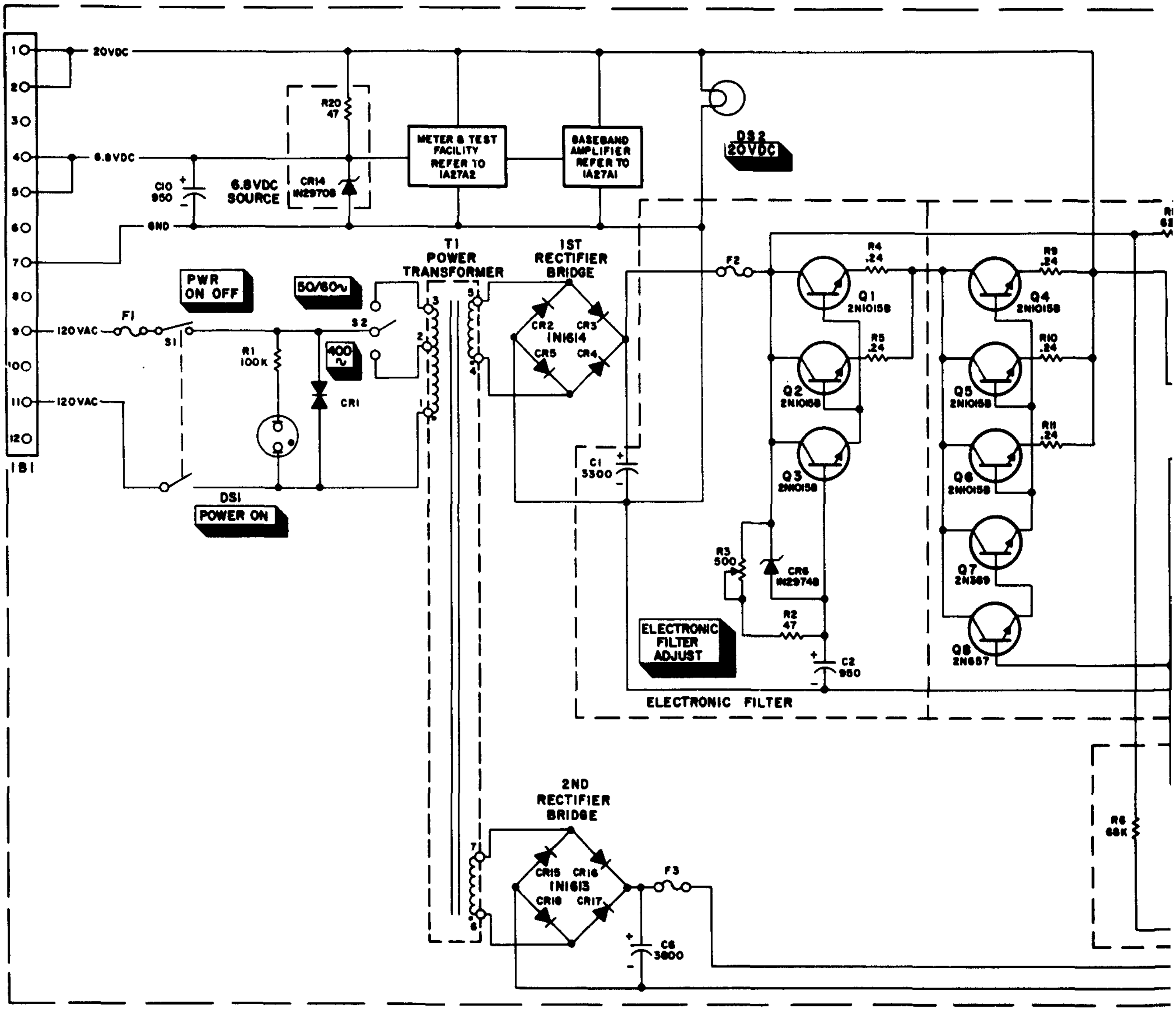
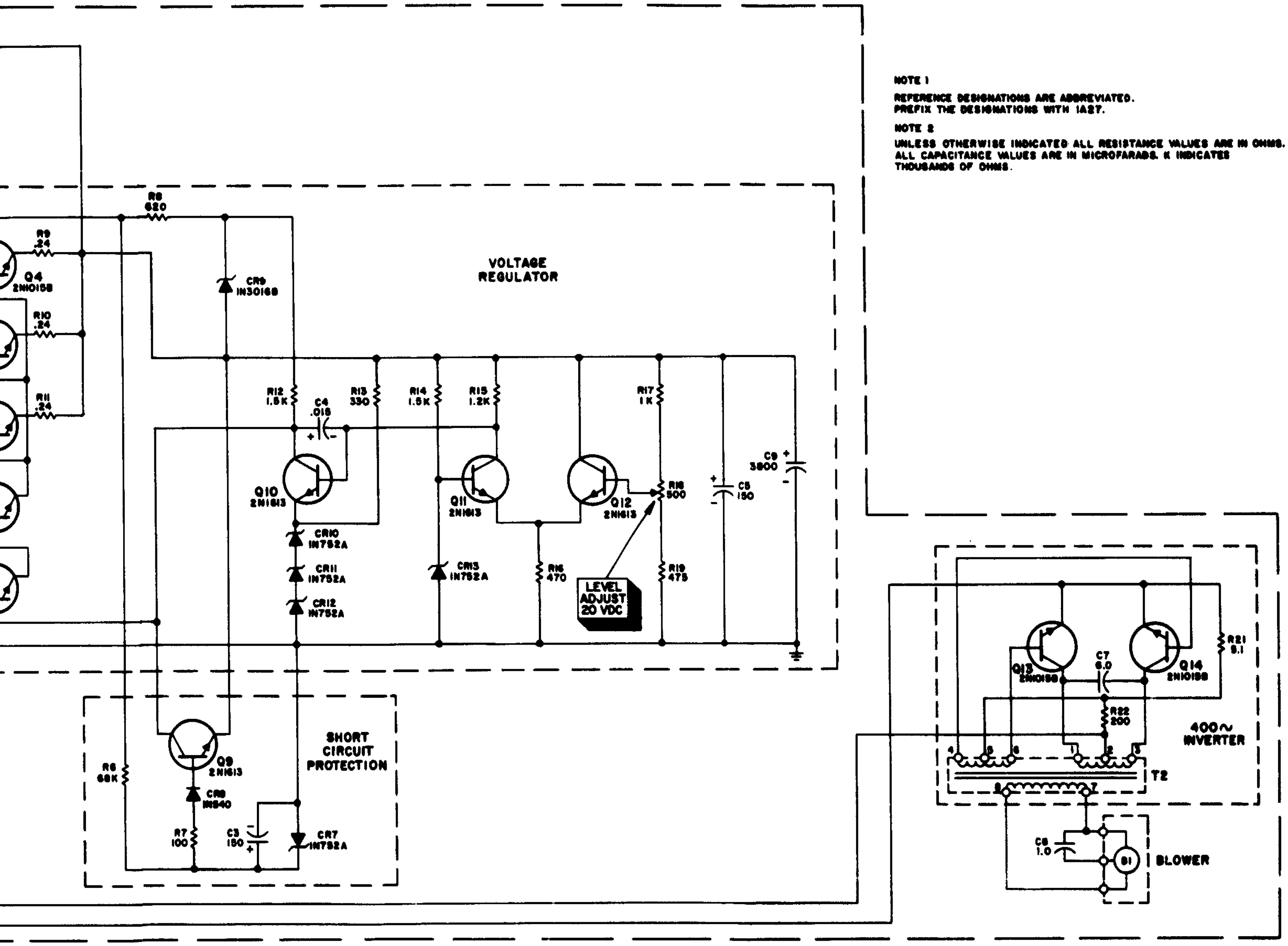


Figure 71 Baseband Amplifier Noise to Noise


20VDC
6.8VDC
GND
120VAC
120VAC
1B1
R20
47
CR14
IN2970B
C10
950
6.8VDC
SOURCE
METER & TEST
FACILITY
REFER TO
1A27A2
BASEBAND
AMPLIFIER
REFER TO
1A27A1
DS2
20VDC
PWR
ON OFF
50/60~
400~
F1
S1
S2
R1
100K
CR1
DS1
POWER ON
T1
POWER
TRANSFORMER
1ST
RECTIFIER
BRIDGE
CR2
CR3
CR5
CR4
IN1614
C1
3300
F2
R4
.24
R5
.24
Q1
2N1015B
Q2
2N1015B
Q3
2N1015B
R9
.24
R10
.24
R11
.24
Q4
2N1015B
Q5
2N1015B
Q6
2N1015B
Q7
2N389
Q8
2N657
R3
500
CR6
IN2974B
R2
47
C2
950
ELECTRONIC
FILTER
ADJUST
ELECTRONIC FILTER
2ND
RECTIFIER
BRIDGE
CR15
CR16
IN1613
CR18
CR17
F3
C6
3600
R6
68K

1



NOTE 1

REFERENCE DESIGNATIONS ARE ABBREVIATED.
PREFIX THE DESIGNATIONS WITH 1A27.

NOTE 2

UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS.
ALL CAPACITANCE VALUES ARE IN MICROFARADS. K INDICATES
THOUSANDS OF OHMS.

Figure 72 Multiplex Power Supply, Schematic Diagram

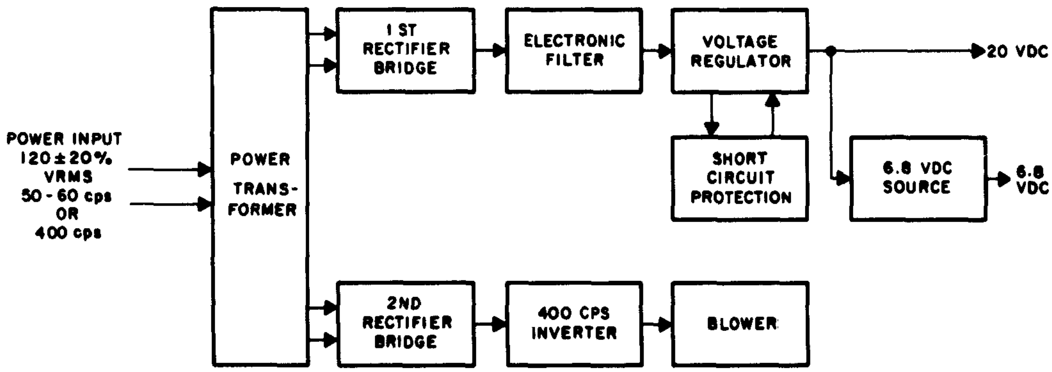


Figure 73 Multiplex Power Supply, Block Diagram

Figure 74   Radio Unit

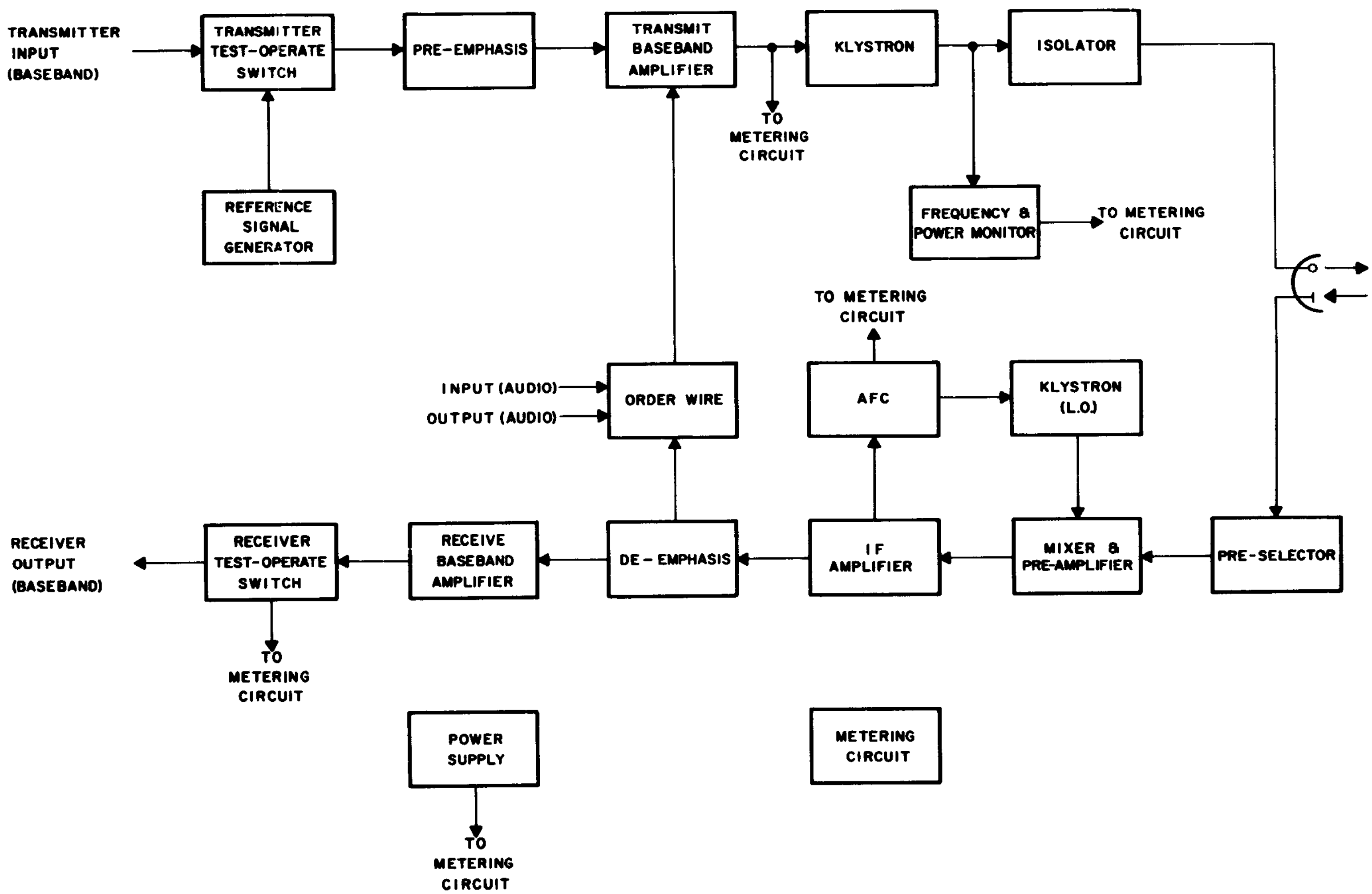


Figure 75 Radio Unit, Block Diagram

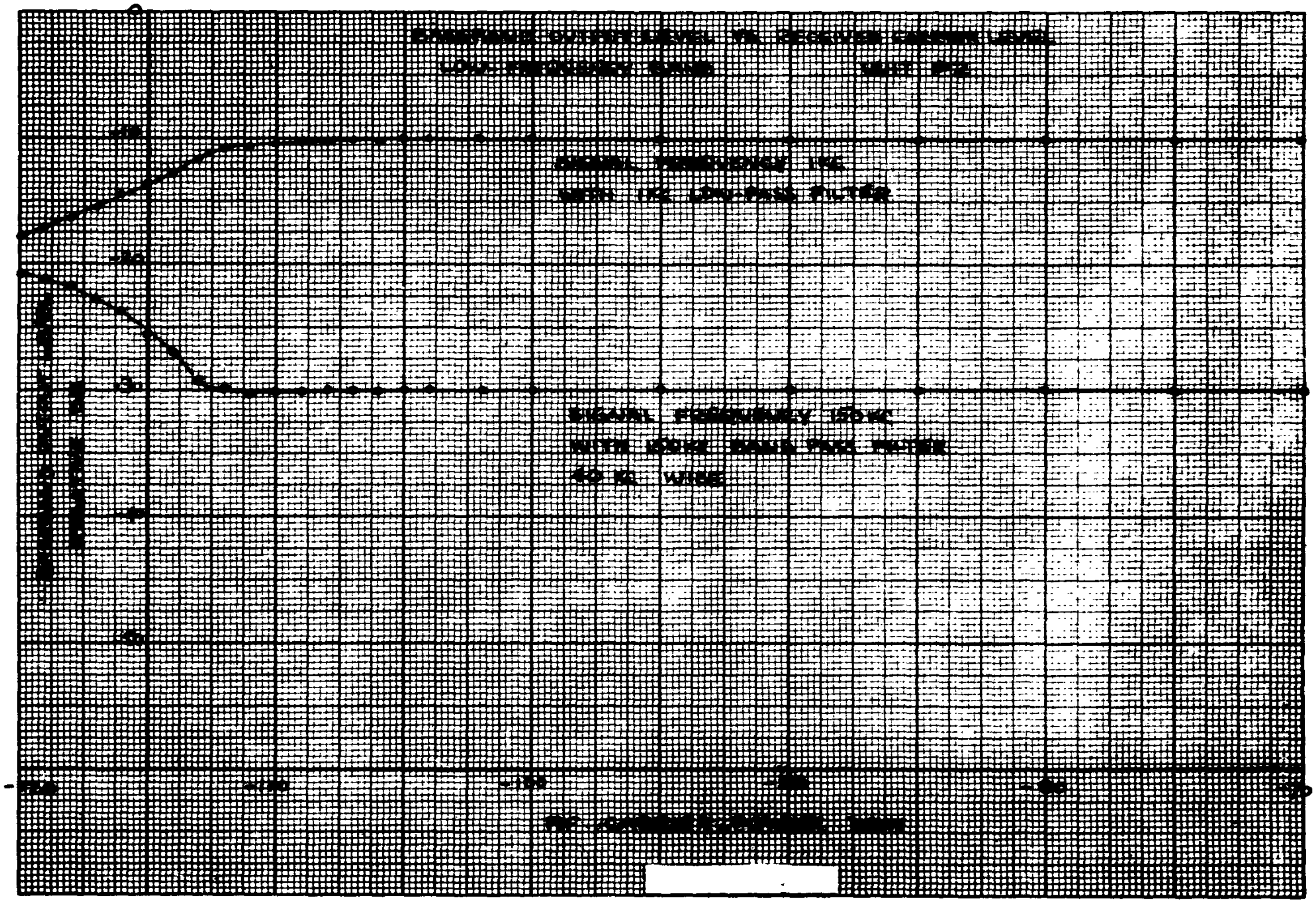


Figure 76 Baseband Output Level Vs. Received Carrier Level, Unit No. 2

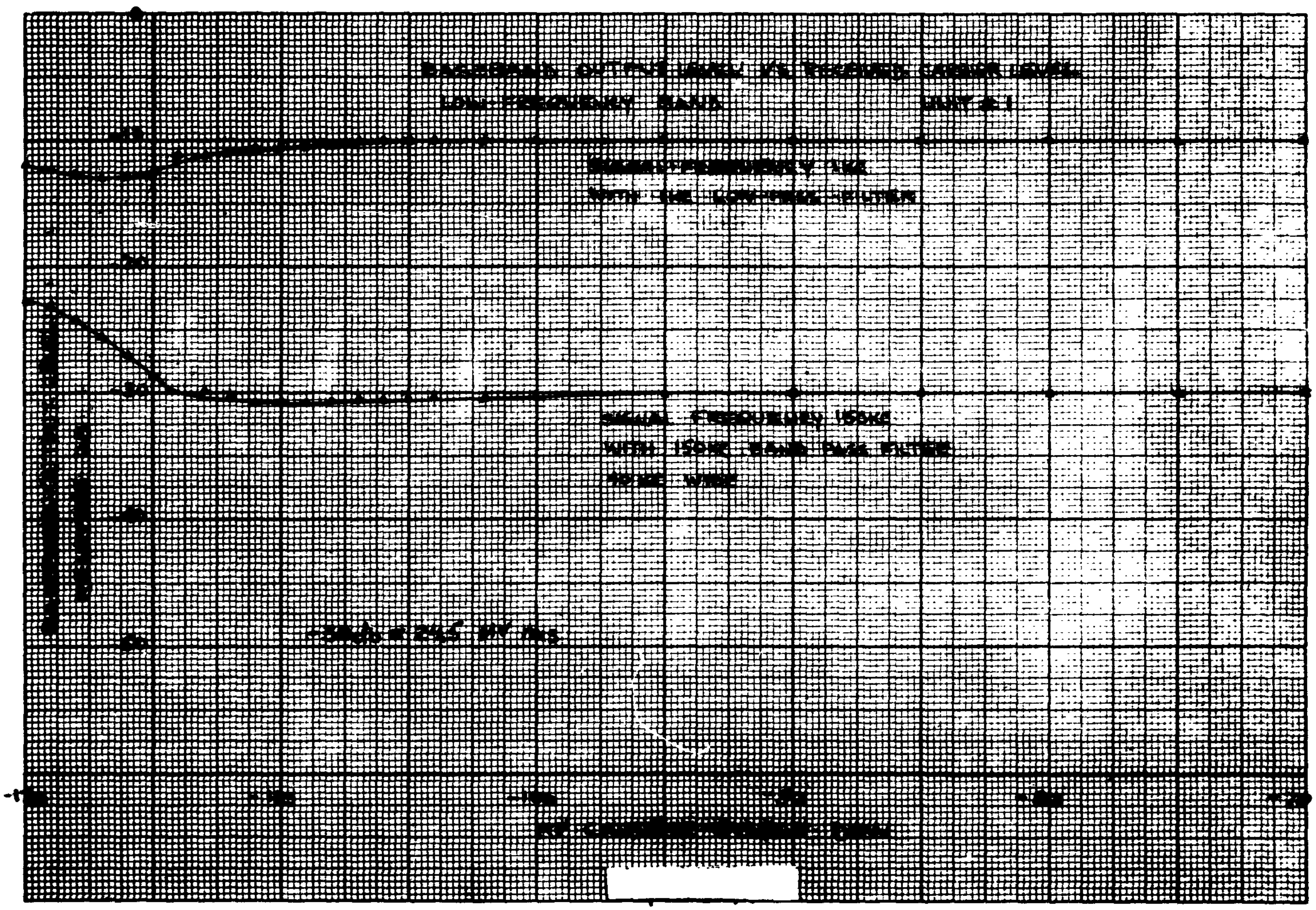
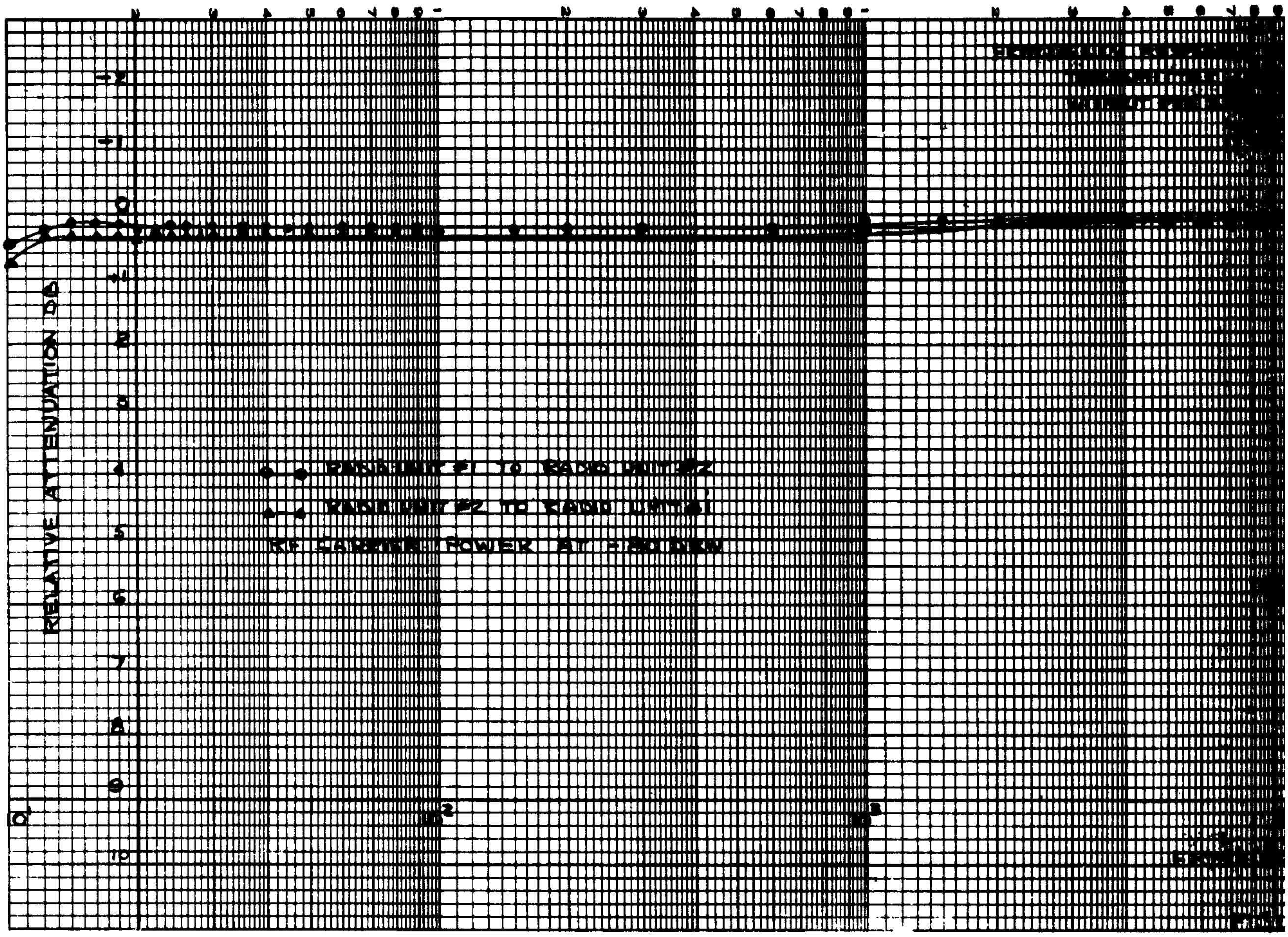


Figure 77 Baseband Output Level Vs. Received Carrier Level, Unit No. 1


RELATIVE ATTENUATION DB
RADIO UNIT #1 TO RADIO UNIT #2
RADIO UNIT #2 TO RADIO UNIT #1
RF CARRIER POWER AT
10
10^2
10^3

1

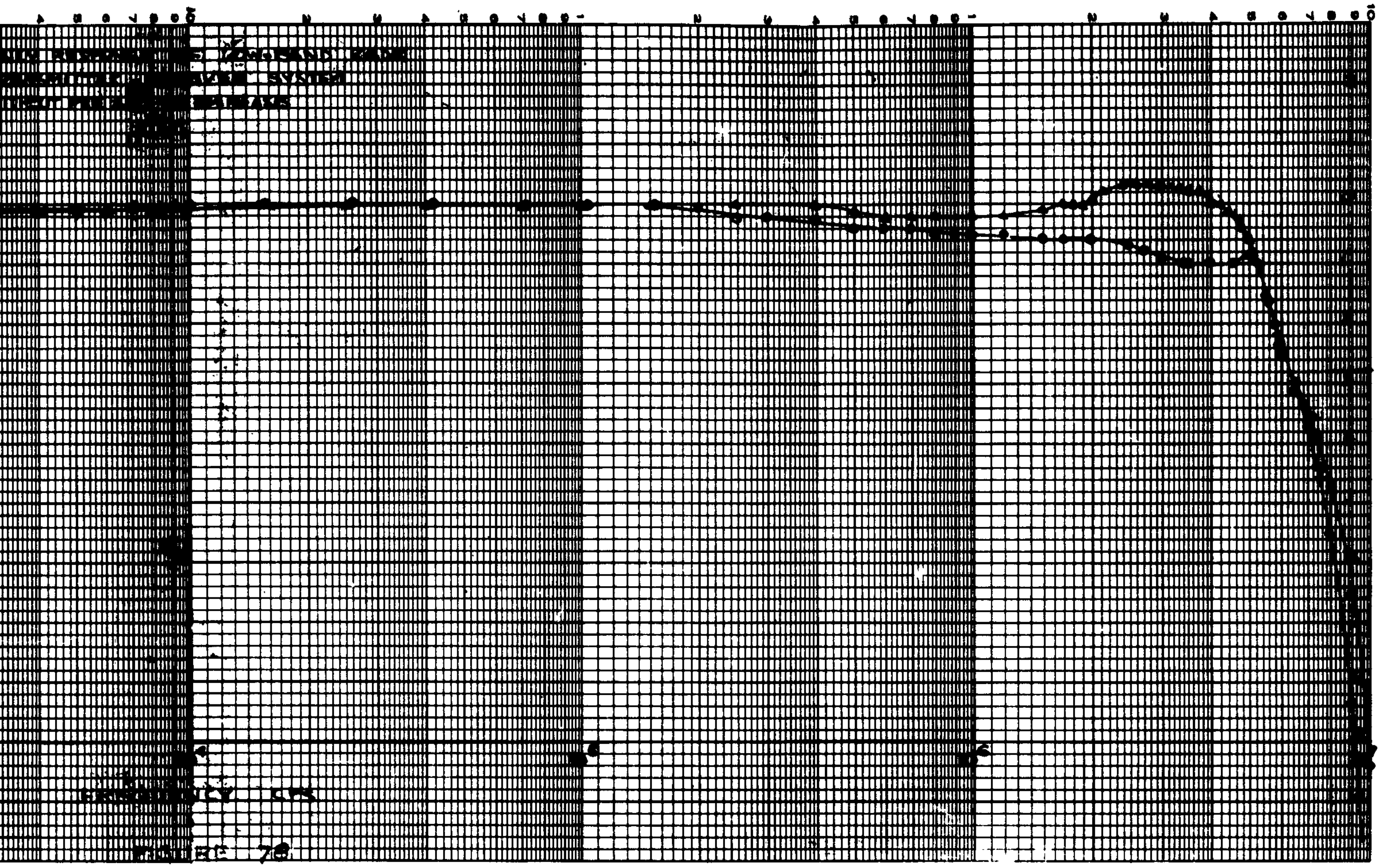
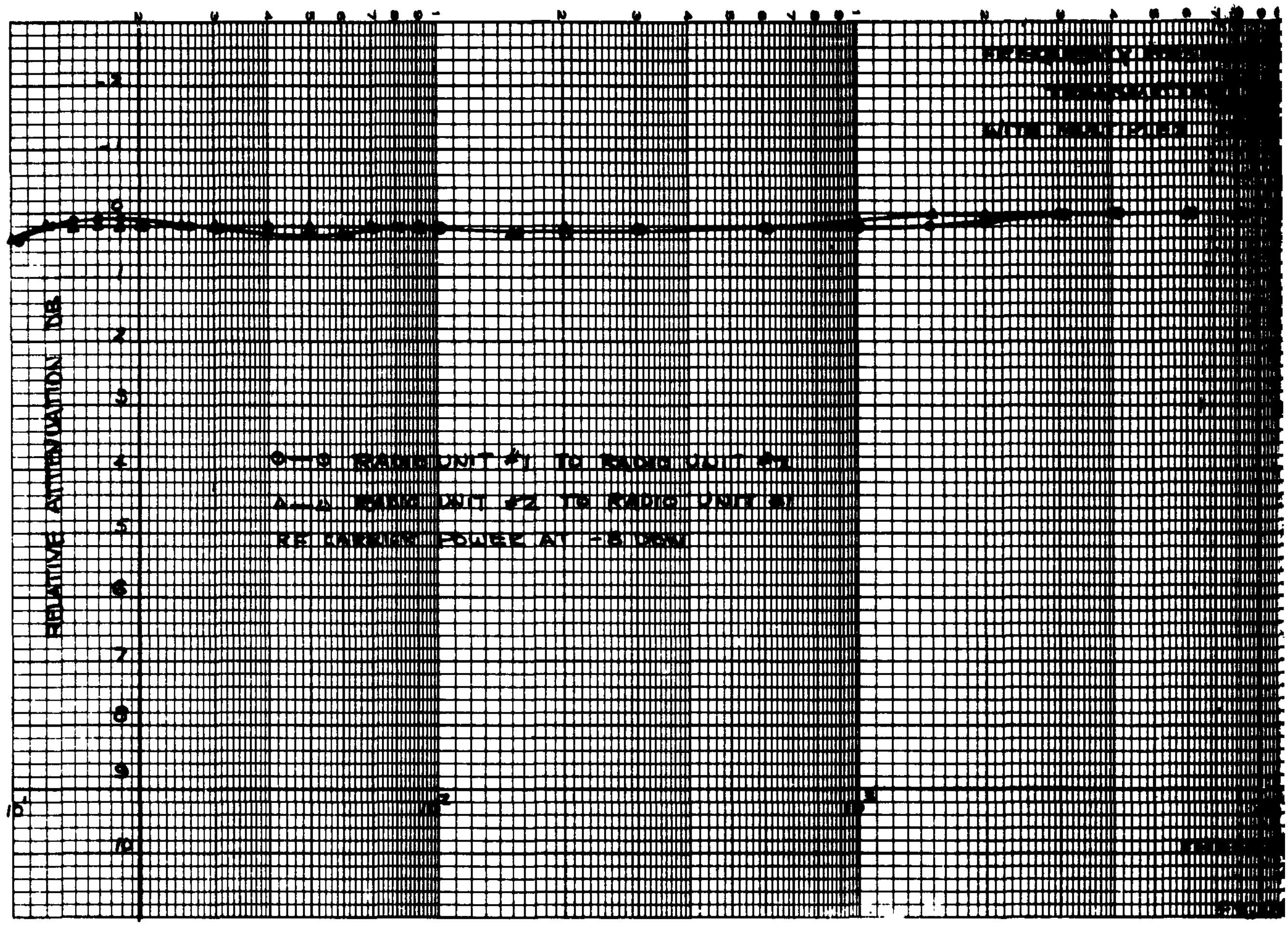


Figure 78 Frequency Response of Low-Band Radio Without Pre- and De-emphasis



1

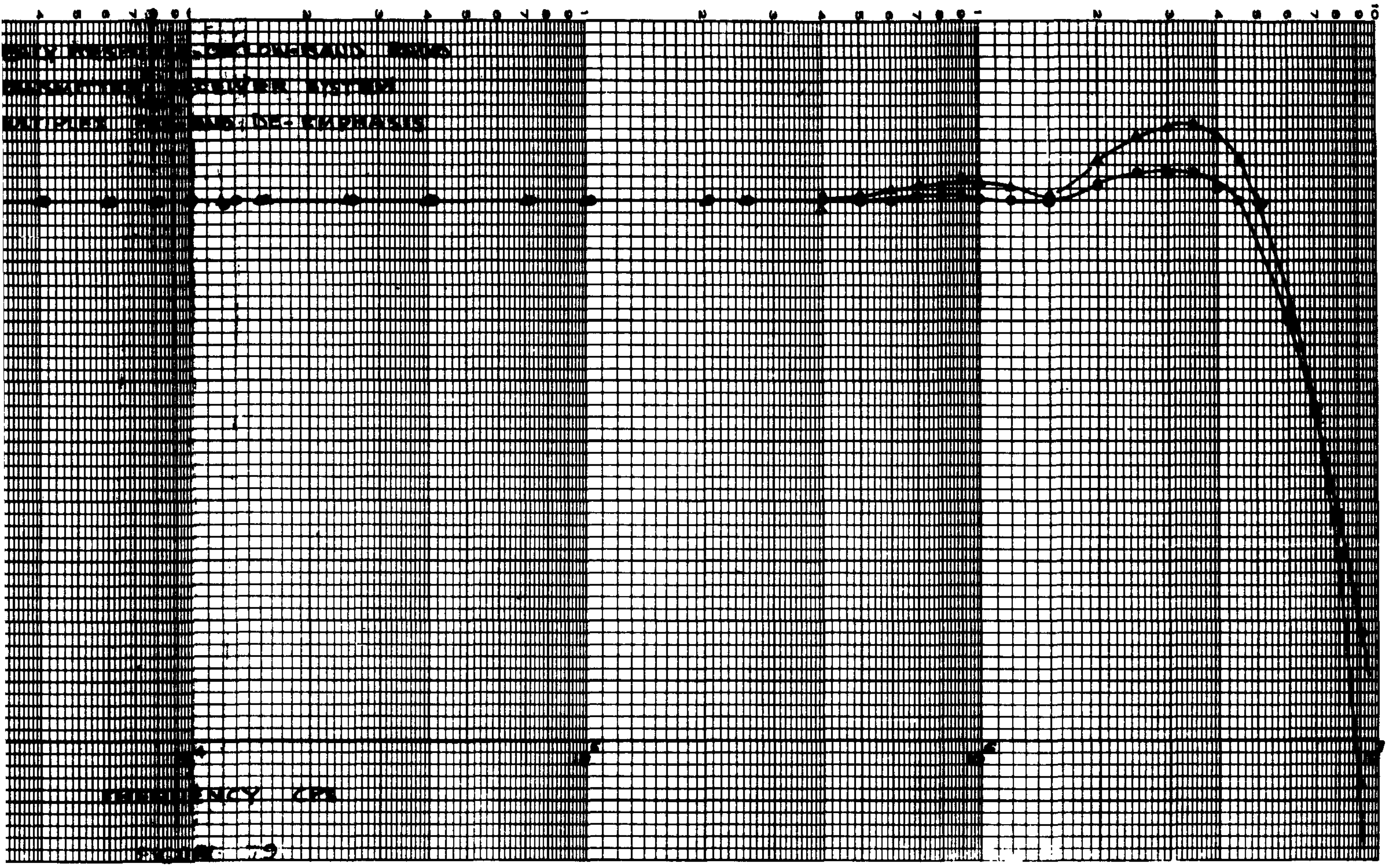
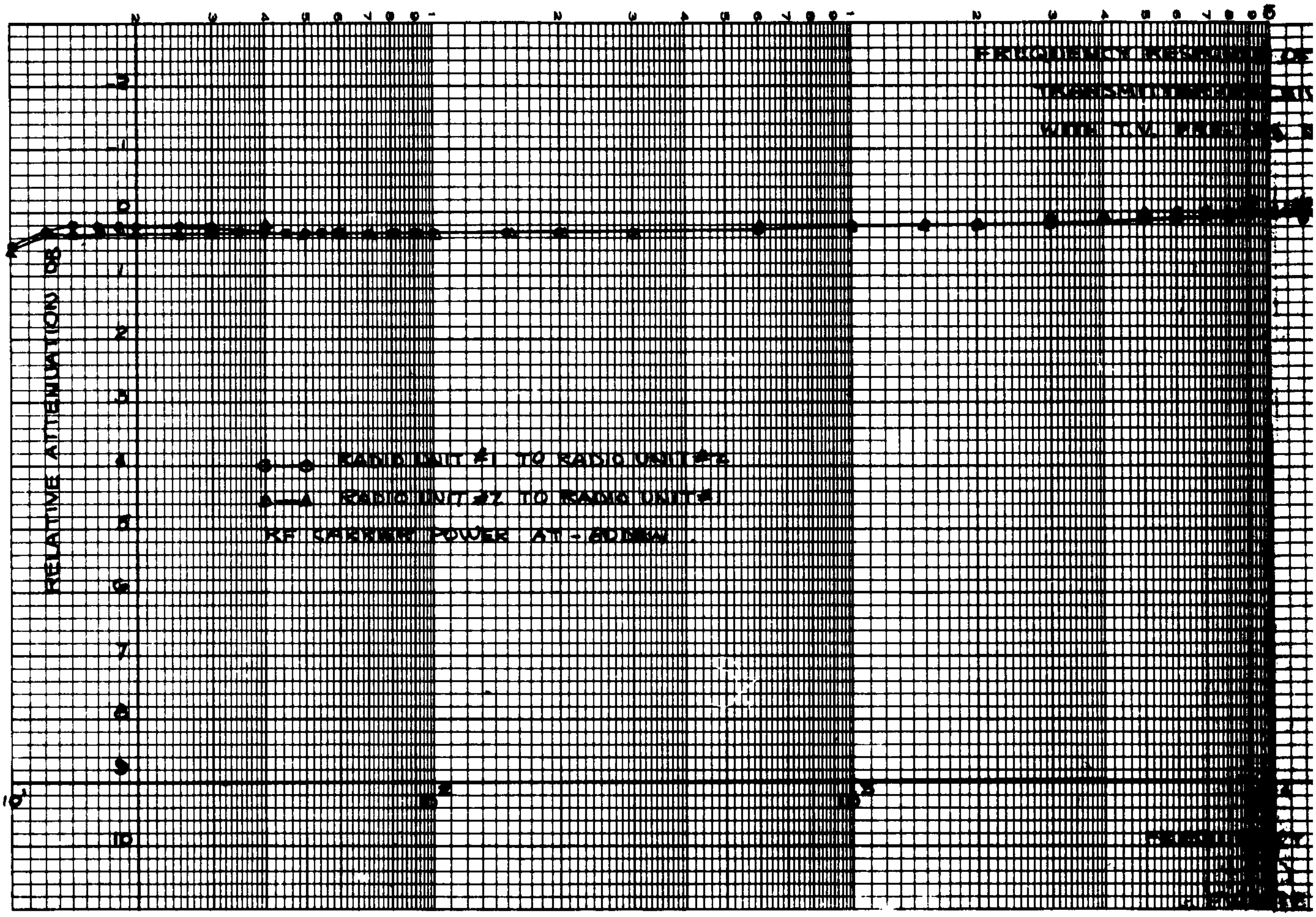


Figure 79 Frequency Response of Low-Band Radio With Multiplex Pre- and De-emphasis


RELATIVE ATTENUATION DB
RADIO UNIT #1 TO RADIO UNIT #2
RADIO UNIT #2 TO RADIO UNIT #1
RF CARRIER POWER AT -80 DBM

1

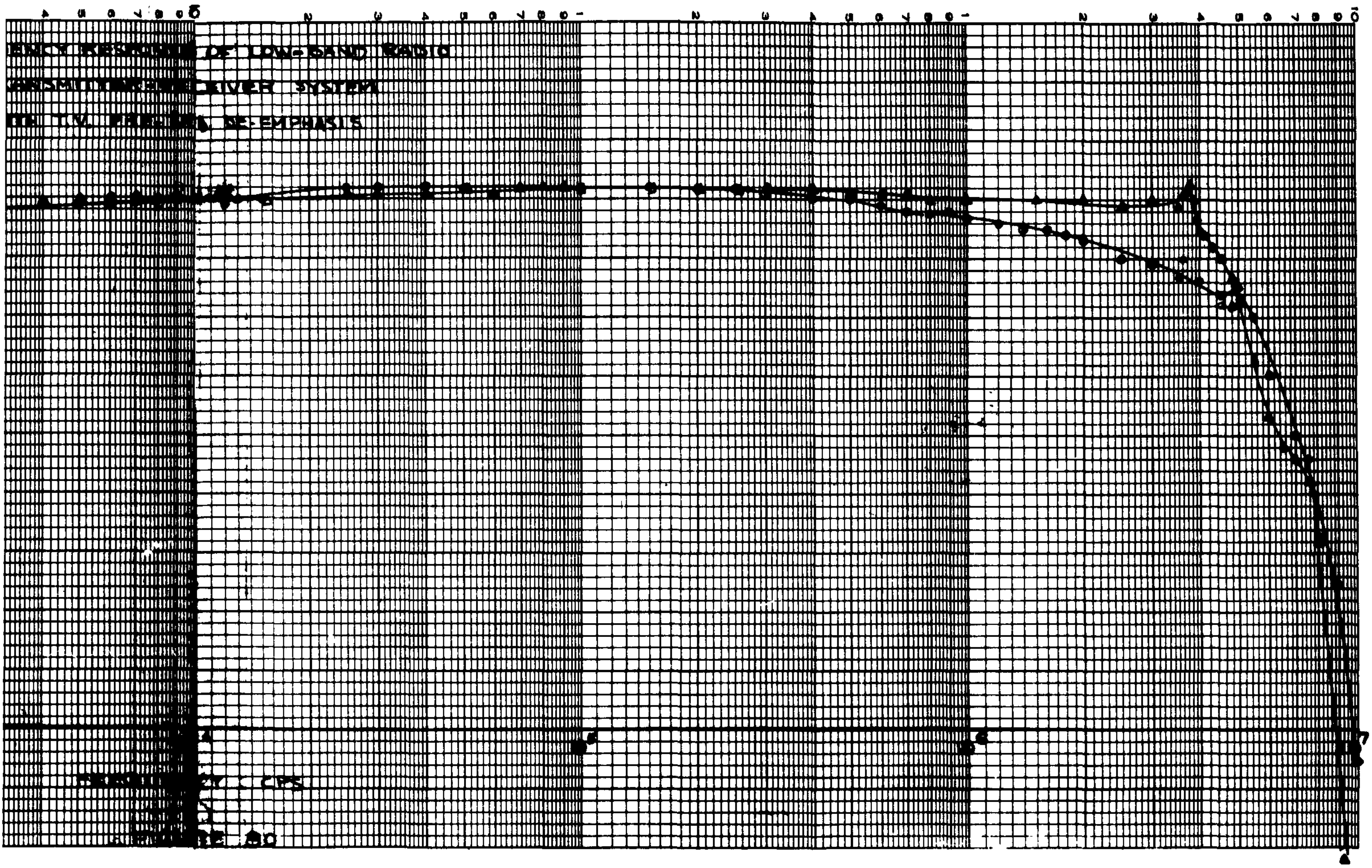
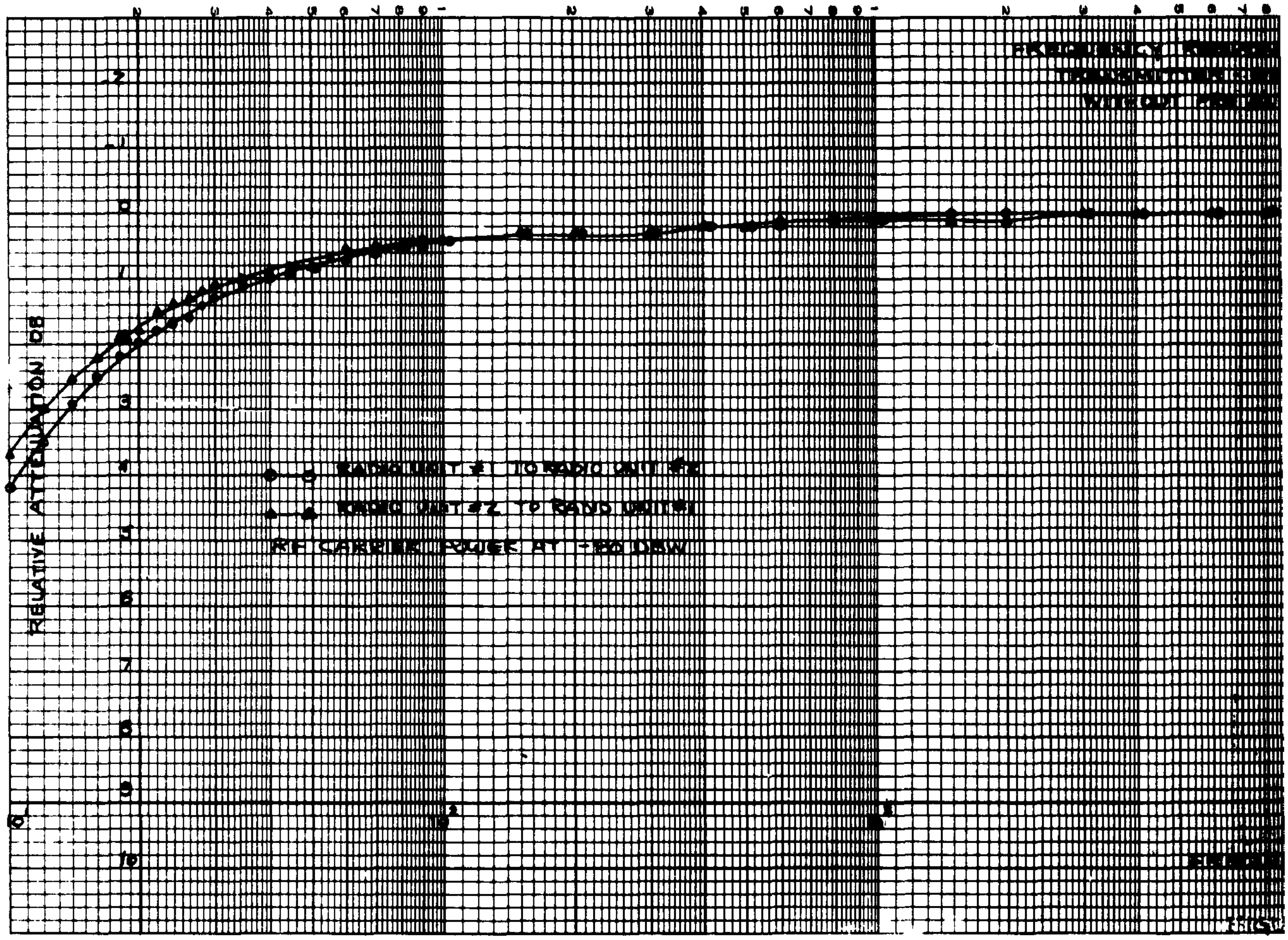


Figure 80 Frequency Response of Low-Band Radio With TV Pre- and De-emphasis



1

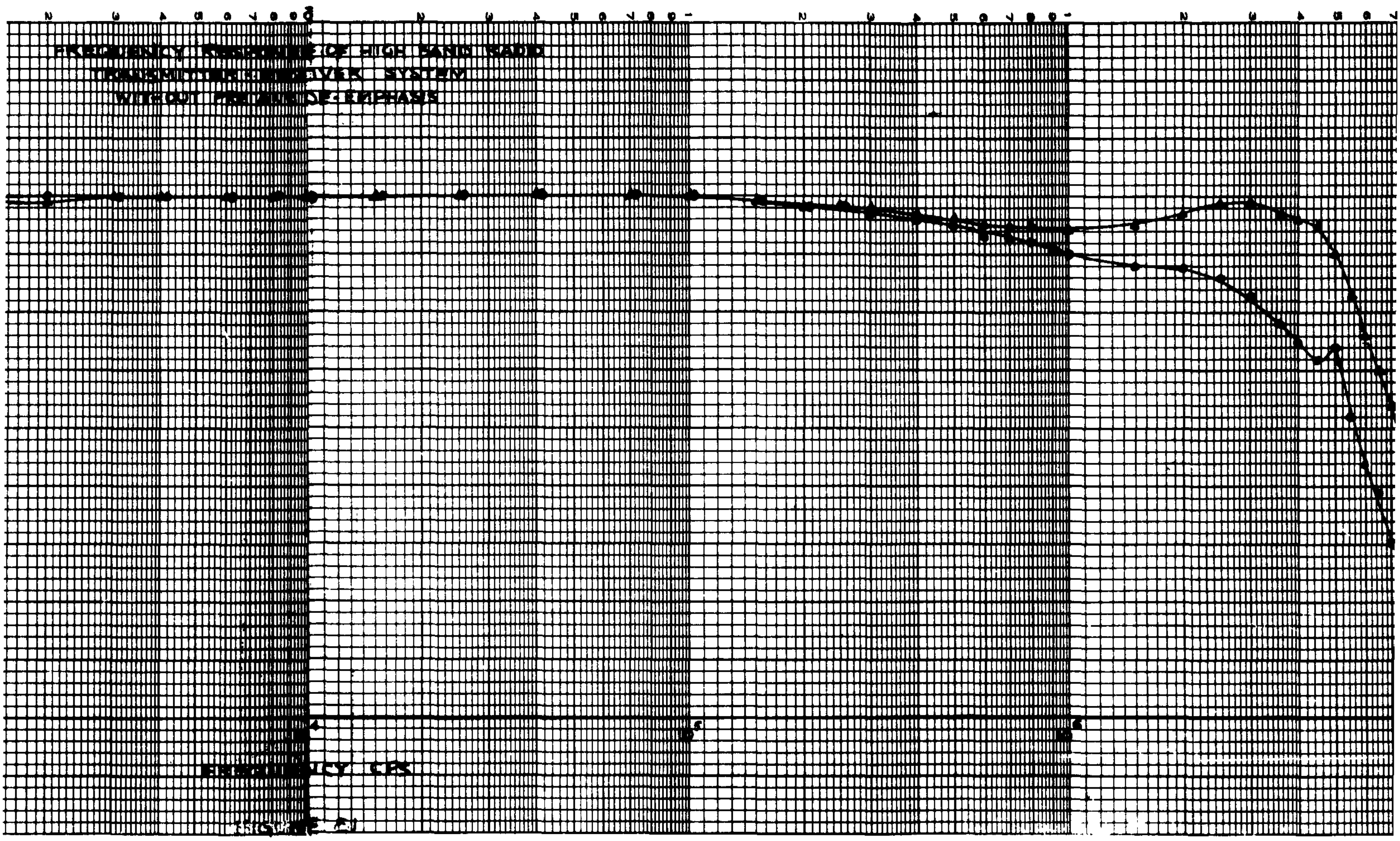


Figure 81 Frequency Response of High-Band Radio Without and De-emphasis

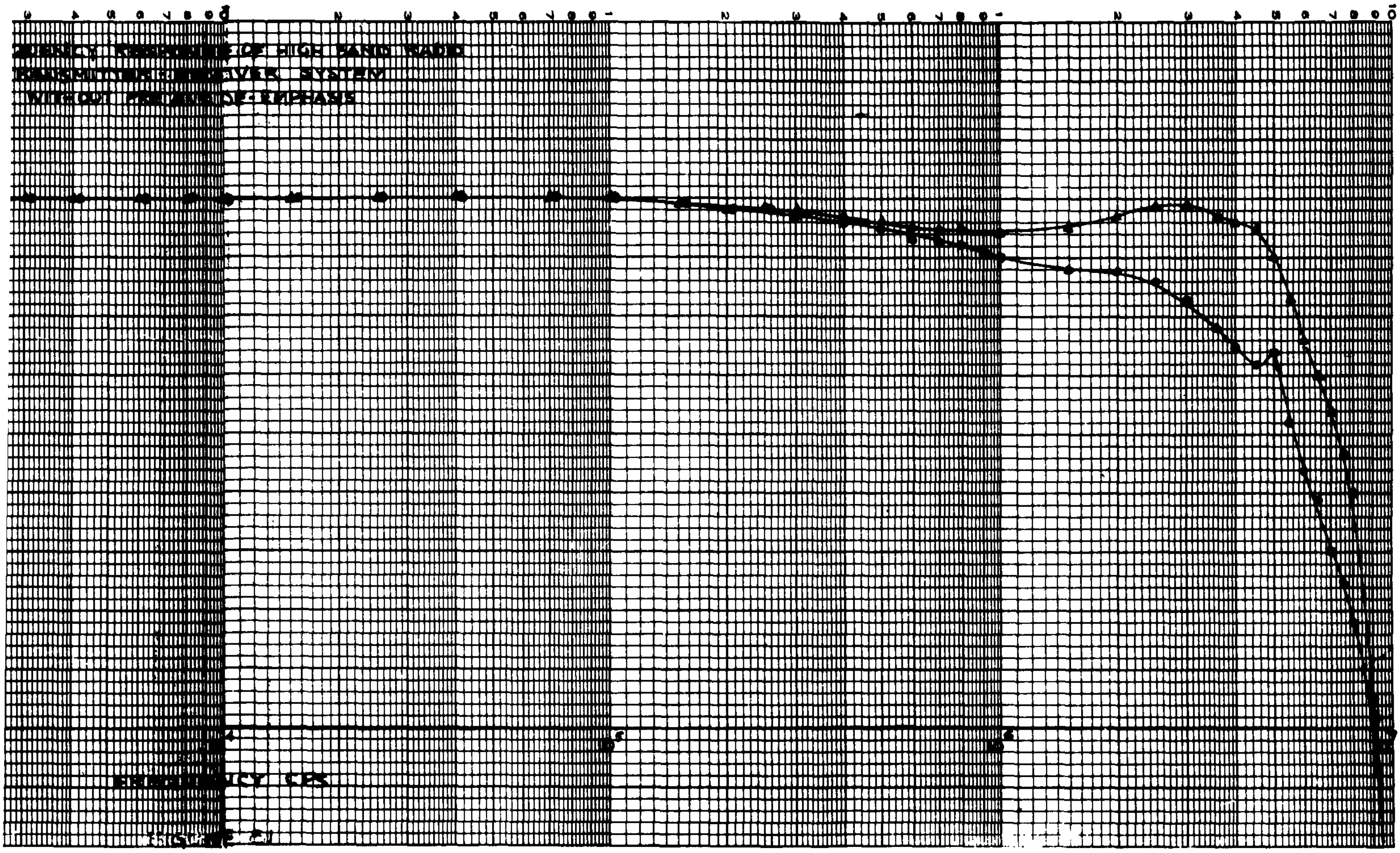


Figure 81 Frequency Response of High-Band Radio Without Pre- and De-emphasis

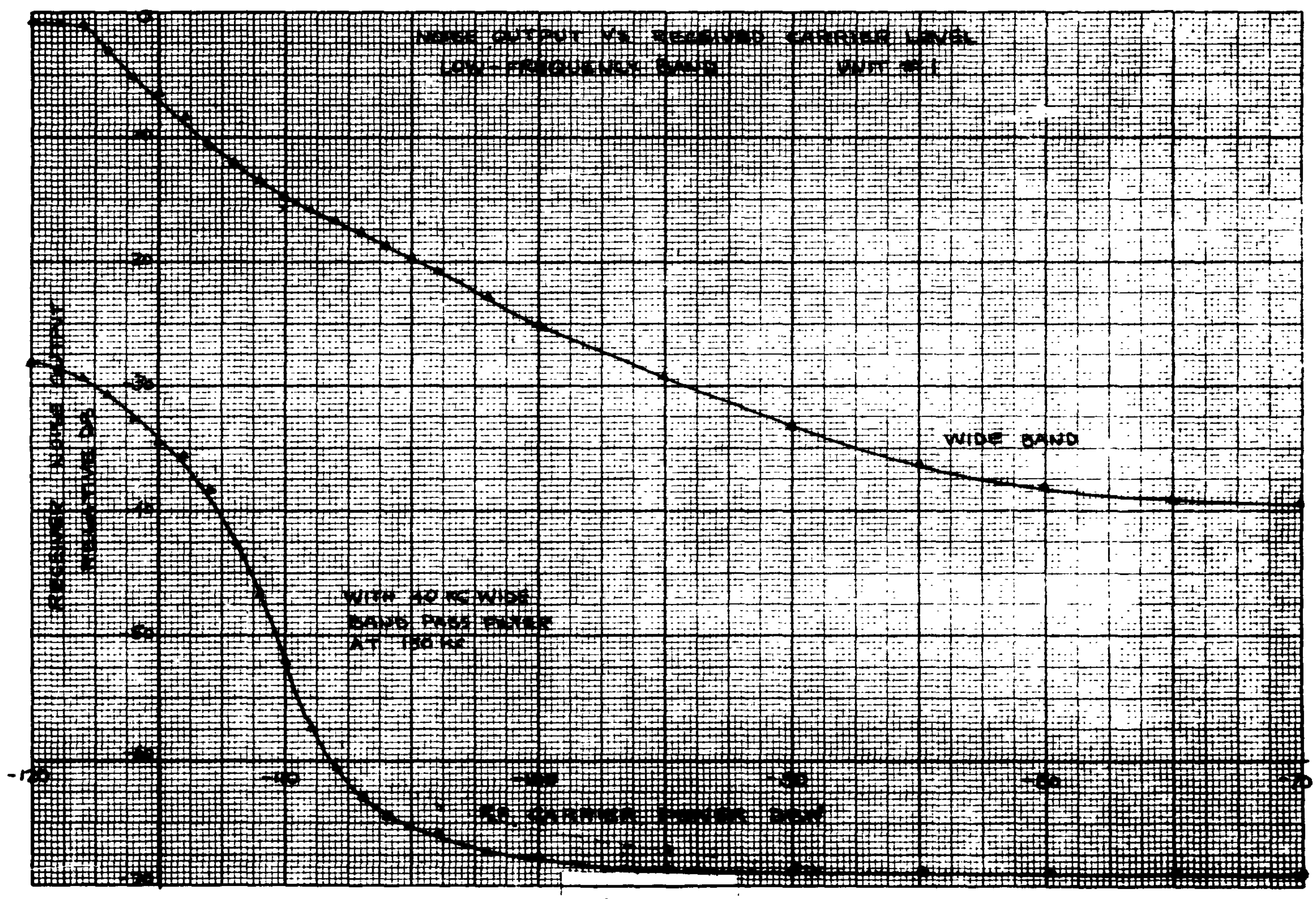


Figure 82 Noise Output Vs. Received Carrier Level, Low-Frequency Band, Unit No. 1

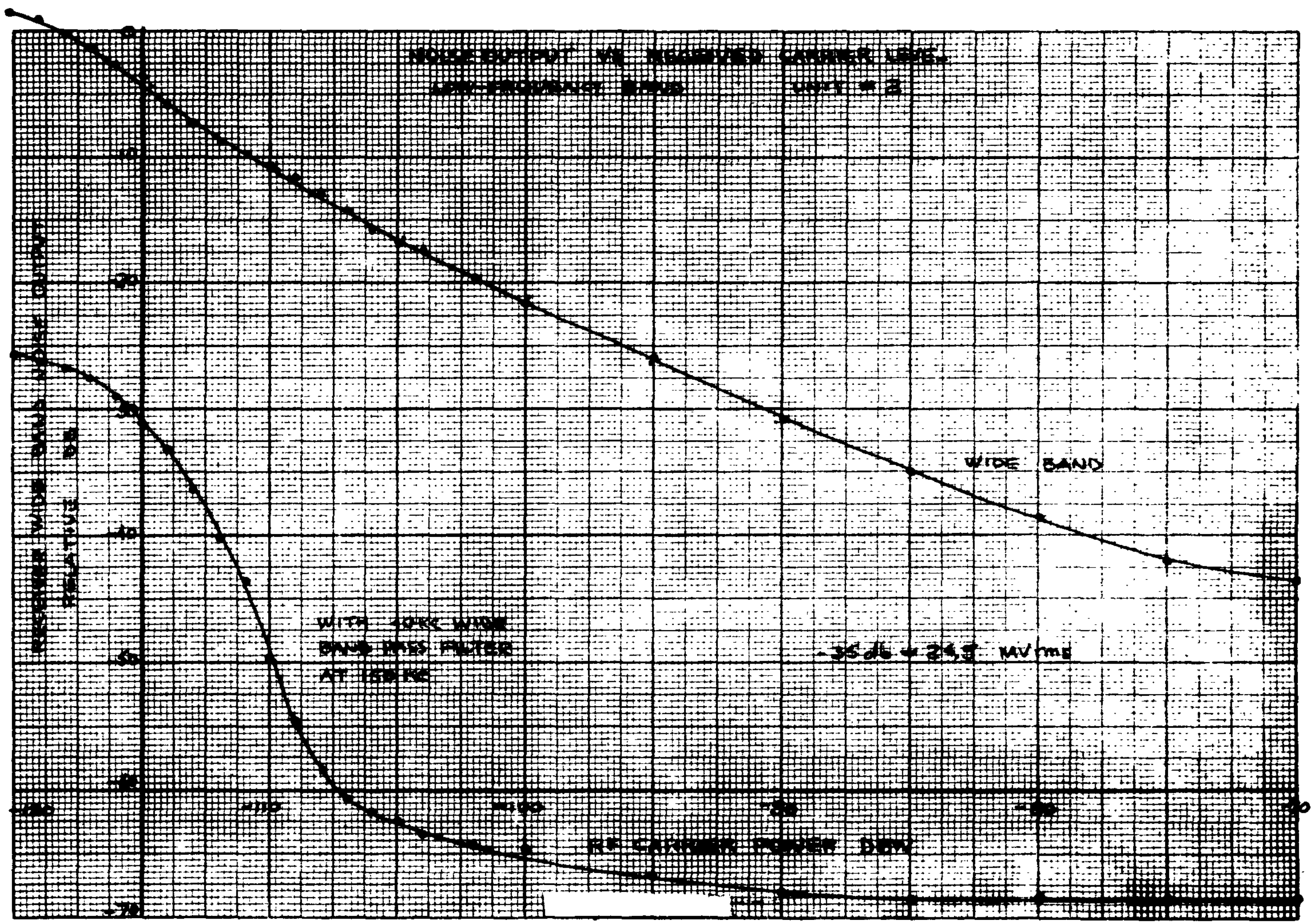


Figure 83 Noise Output Vs. Received Carrier Level, Low-Frequency Band, Unit No. 2

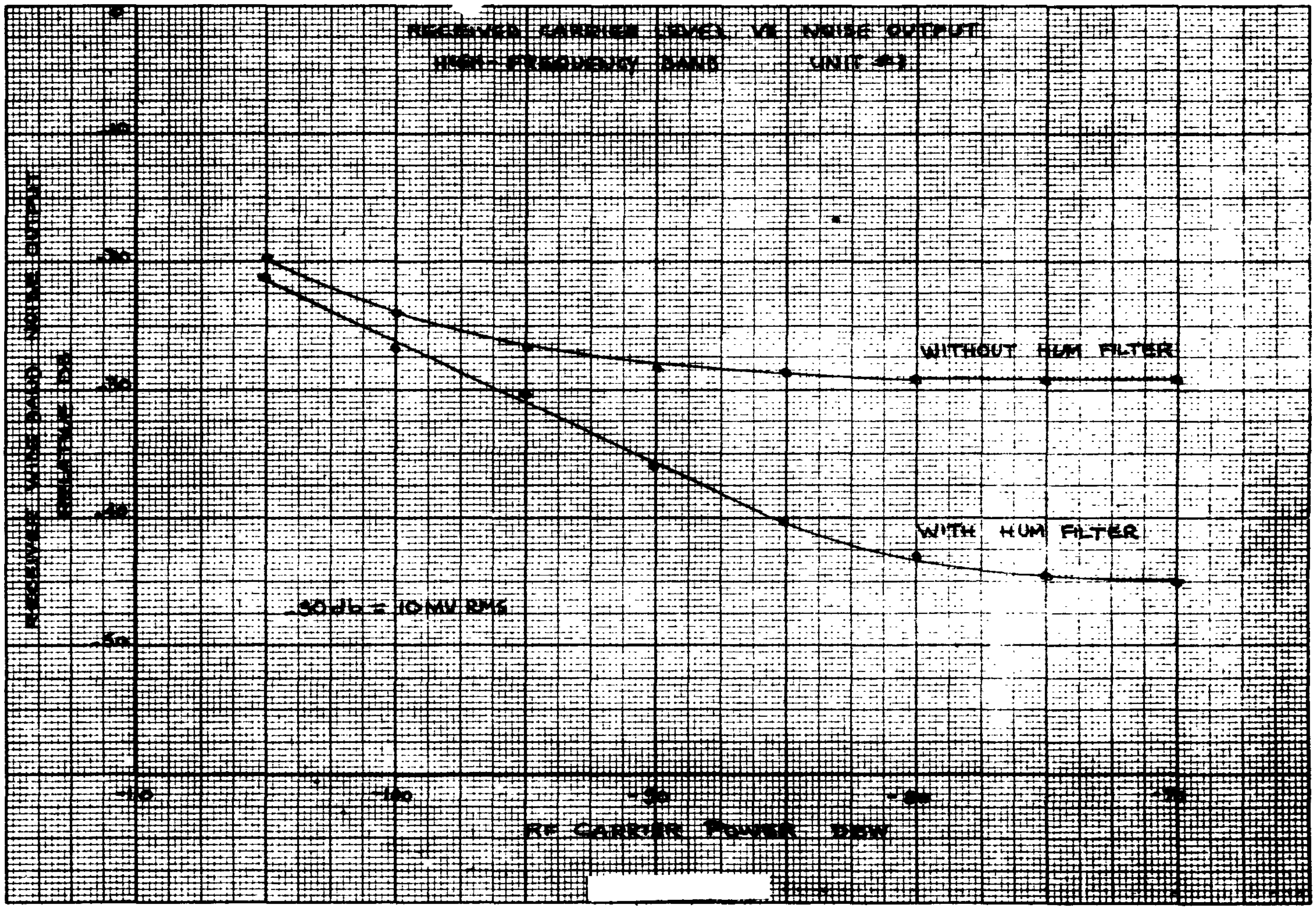


Figure 84 Received Carrier Level Vs. Noise Output, High-Frequency Band, Unit No. 1

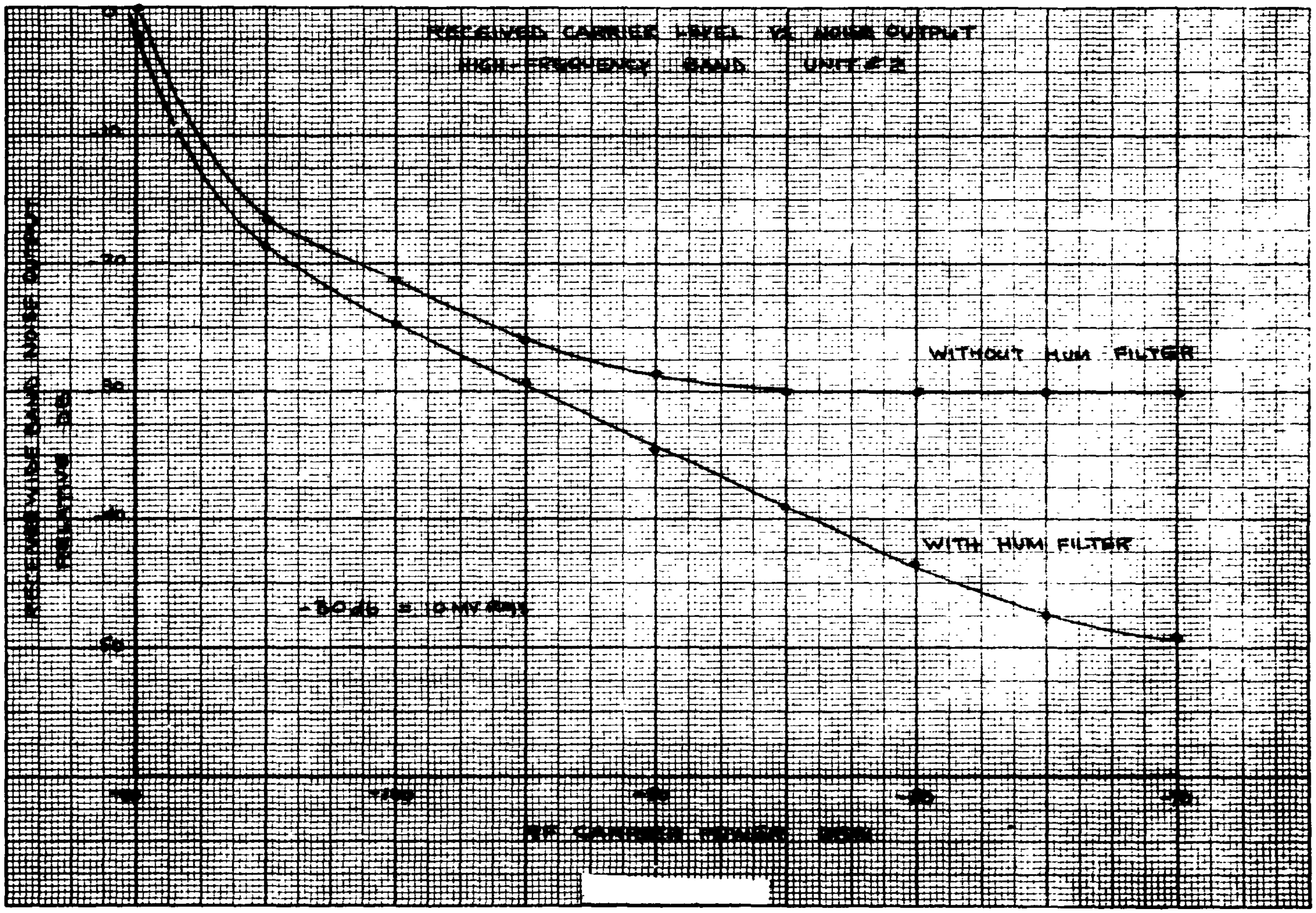
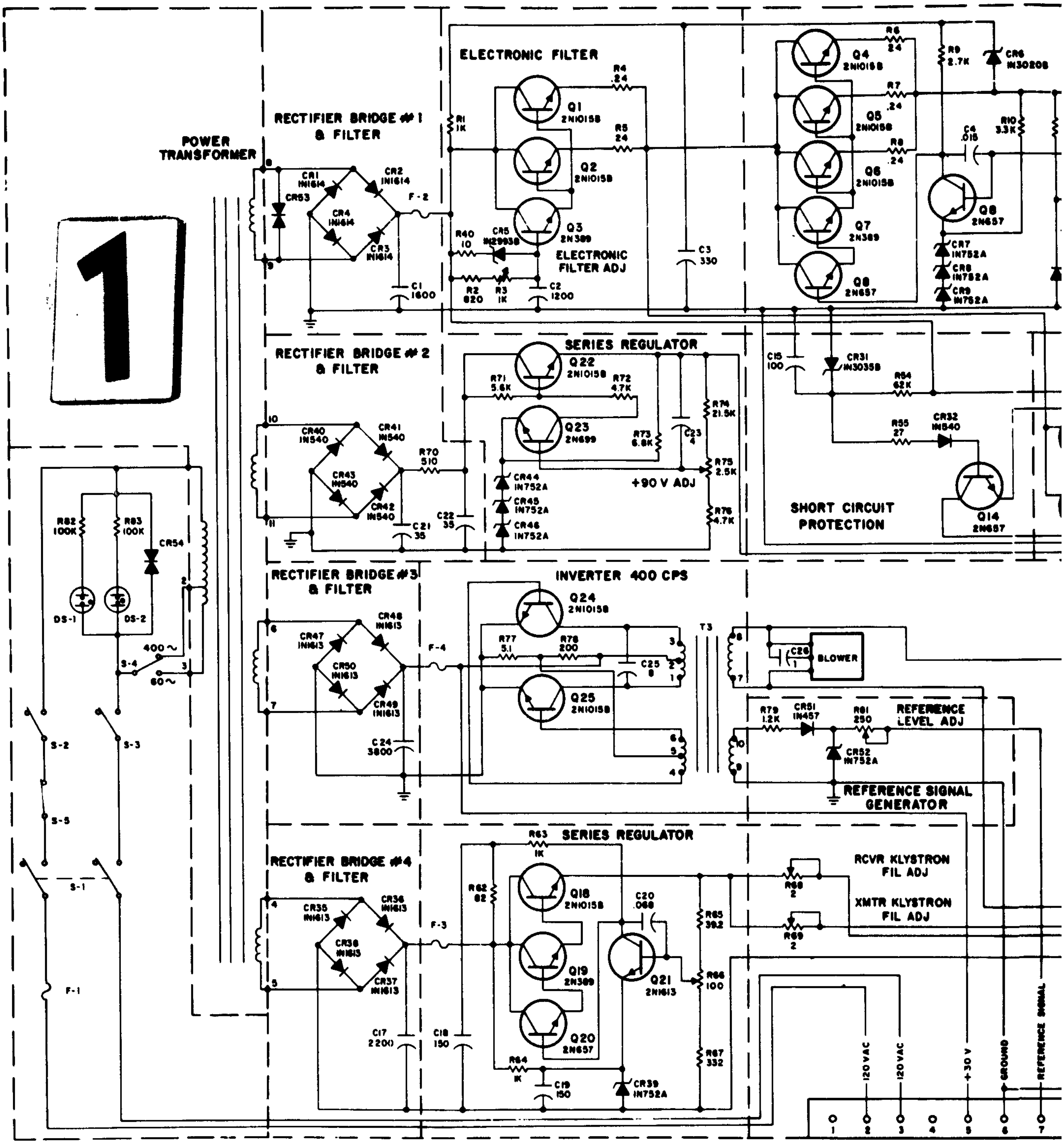


Figure 85 Received Carrier Level Vs. Noise Output, High-Frequency Band, Unit No. 2


1
POWER TRANSFORMER
RECTIFIER BRIDGE #1 & FILTER
ELECTRONIC FILTER
ELECTRONIC FILTER ADJ
RECTIFIER BRIDGE #2 & FILTER
SERIES REGULATOR
+90 V ADJ
SHORT CIRCUIT PROTECTION
RECTIFIER BRIDGE #3 & FILTER
INVERTER 400 CPS
BLOWER
REFERENCE LEVEL ADJ
REFERENCE SIGNAL GENERATOR
RECTIFIER BRIDGE #4 & FILTER
SERIES REGULATOR
RCVR KLYSTRON FIL ADJ
XMTR KLYSTRON FIL ADJ
120 VAC
120 VAC
+30 V
GROUND
REFERENCE SIGNAL
400 ~
60 ~
S-1
S-2
S-3
S-4
S-5
F-1
F-2
F-3
F-4
T3
DS-1
DS-2
Q1 2N1015B
Q2 2N1015B
Q3 2N389
Q4 2N1015B
Q5 2N1015B
Q6 2N1015B
Q7 2N389
Q8 2N657
Q14 2N657
Q18 2N1015B
Q19 2N389
Q20 2N657
Q21 2N1613
Q22 2N1015B
Q23 2N699
Q24 2N1015B
Q25 2N1015B

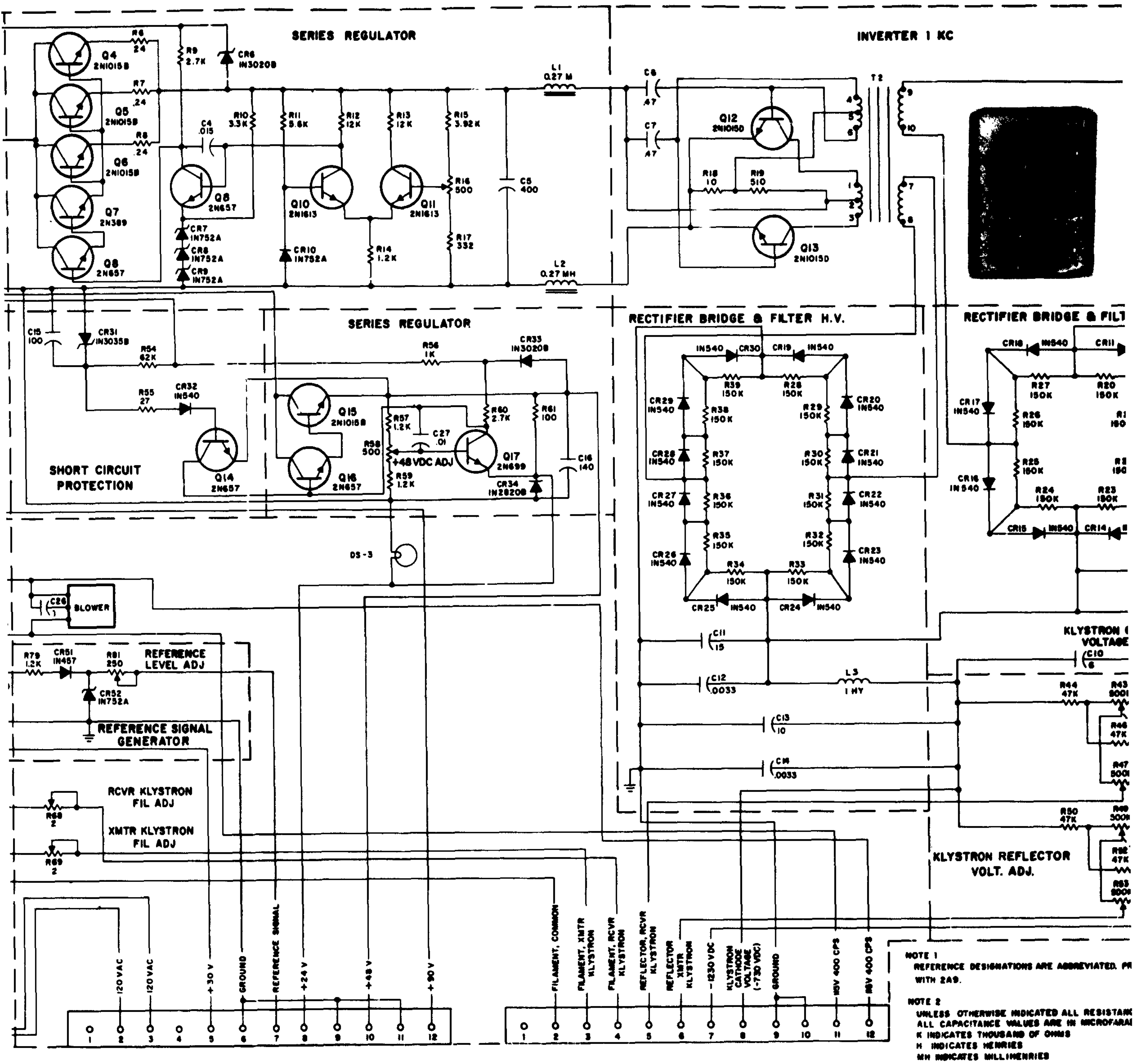
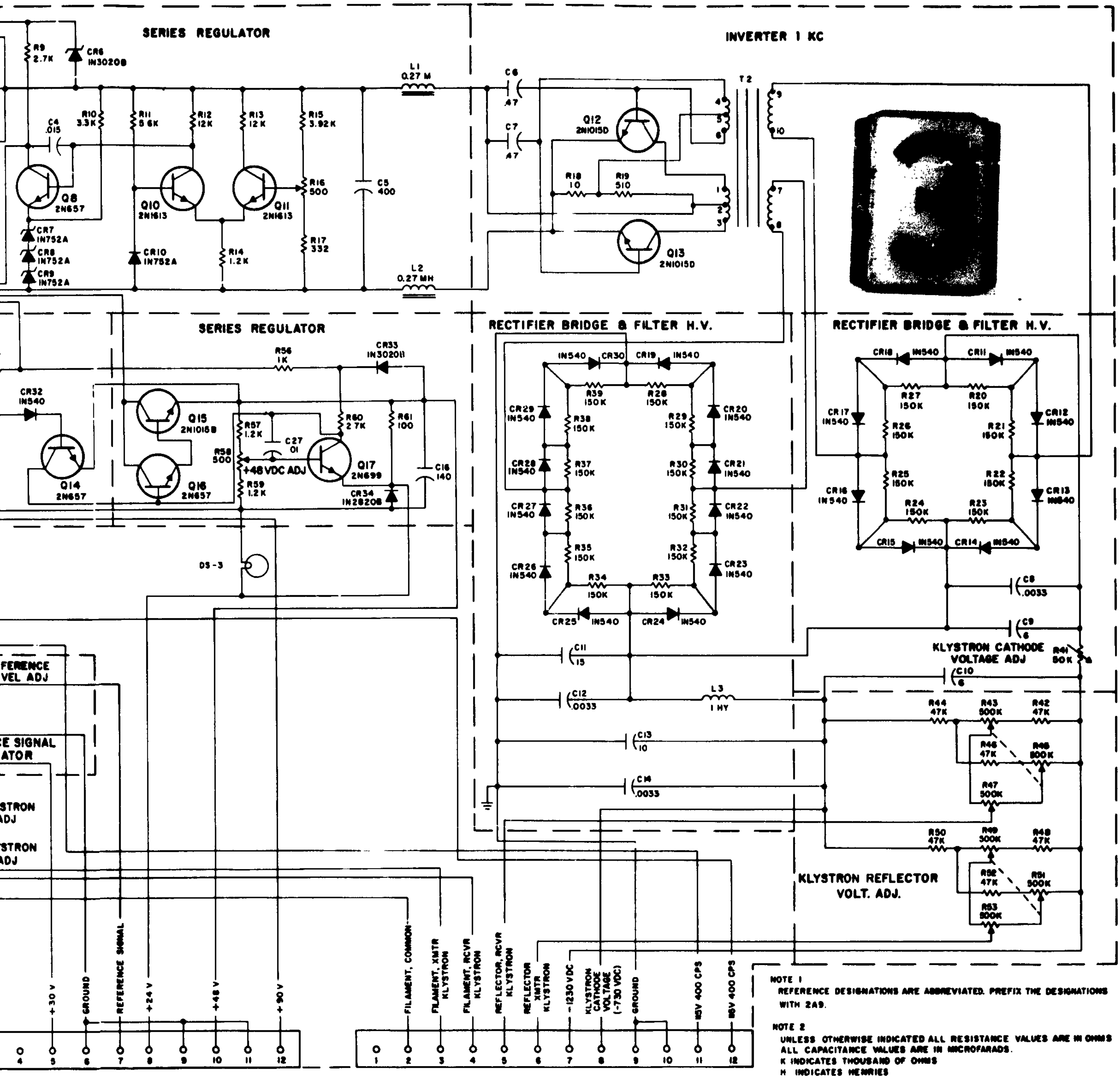


Figure 86 Radio Power Supply, Schemat



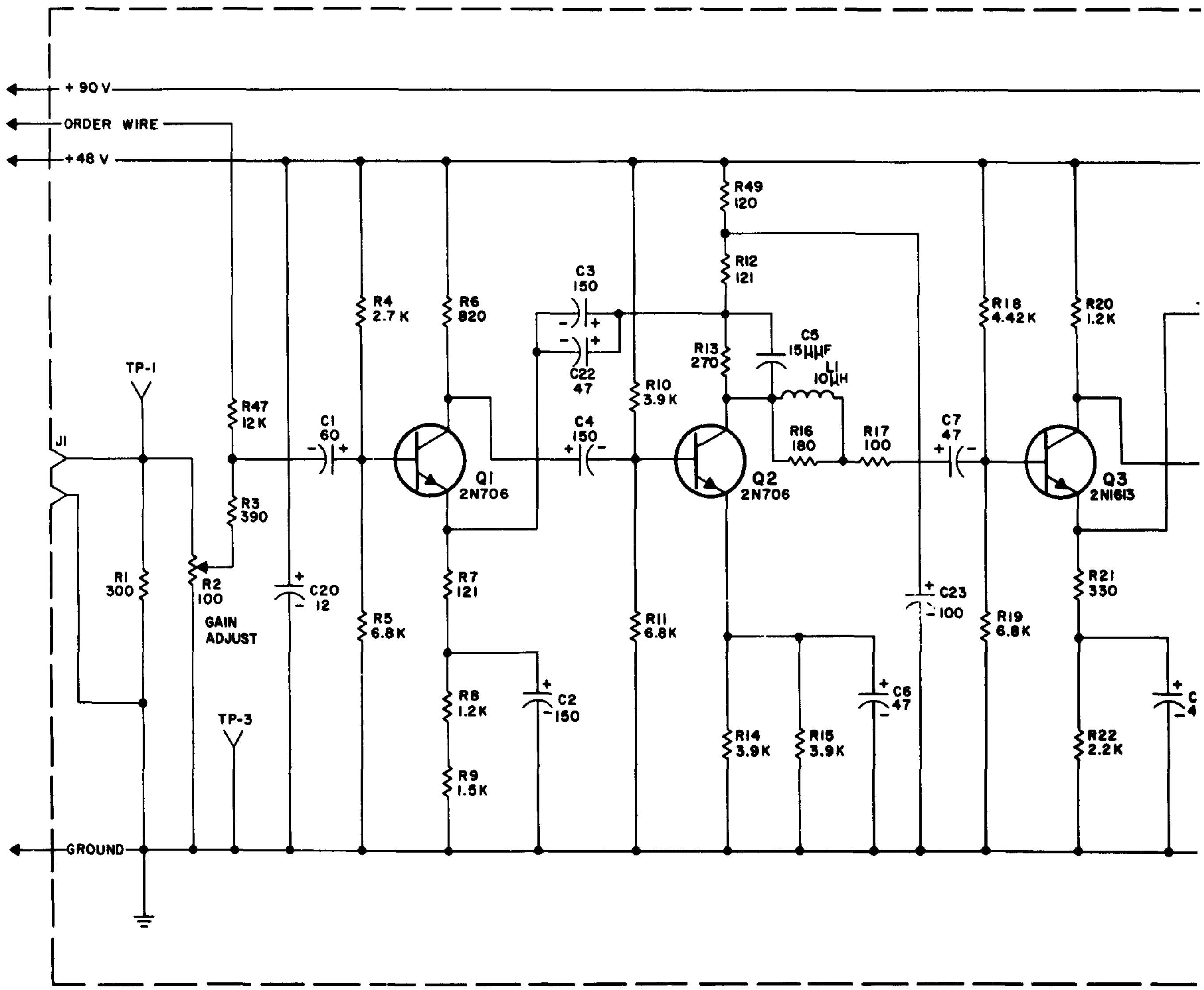
NOTE 1
REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE DESIGNATIONS WITH 2A9.

NOTE 2
UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS
ALL CAPACITANCE VALUES ARE IN MICROFARADS.
K INDICATES THOUSAND OF OHMS
H INDICATES HENRIES
MH INDICATES MILLIHENRIES

Figure 86 Radio Power Supply, Schematic Diagram

NOTE 1
REFERENCE DESIGNATIONS
ARE ABBREVIATED.
PREFIX THE DESIGNATIONS
WITH 2A4.

NOTE 2
UNLESS OTHERWISE INDICATED
ALL RESISTANCE VALUES ARE IN OHMS
ALL CAPACITANCE VALUES ARE IN MICROF
K INDICATES THOUSANDS OF OHMS
μμF INDICATES MICROMICROFARADS
μH INDICATES MICROHENRIES
MEG INDICATES MILLIONS OF OHMS



1

INDICATED
ALUES ARE IN OHMS
VALUES ARE IN MICROFARADS
ANDS OF OHMS
ROMICROFARADS
ROHENRIES
LIONS OF OHMS

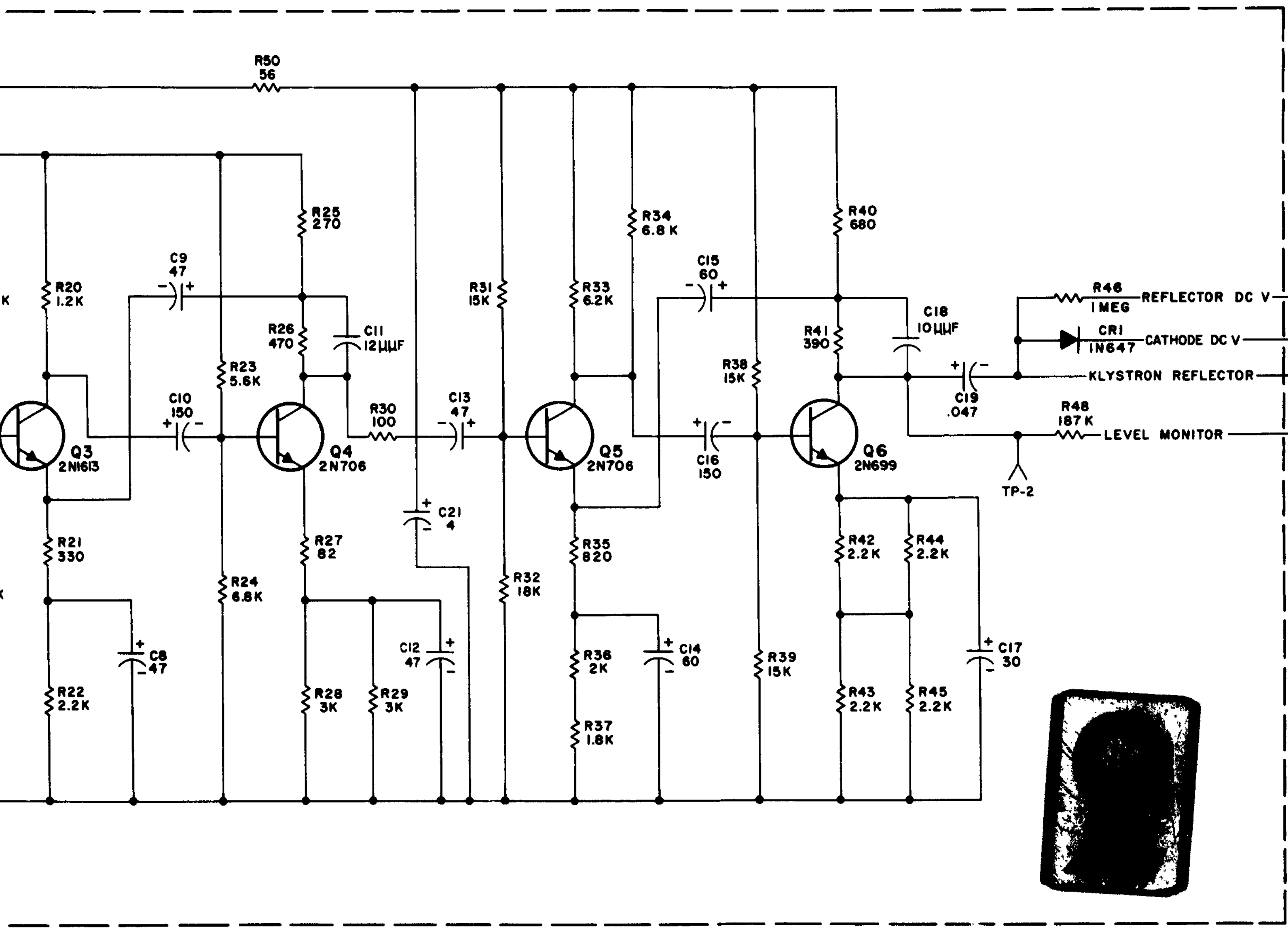


Figure 87 Transmit Baseband Amplifier, Schematic Diagram

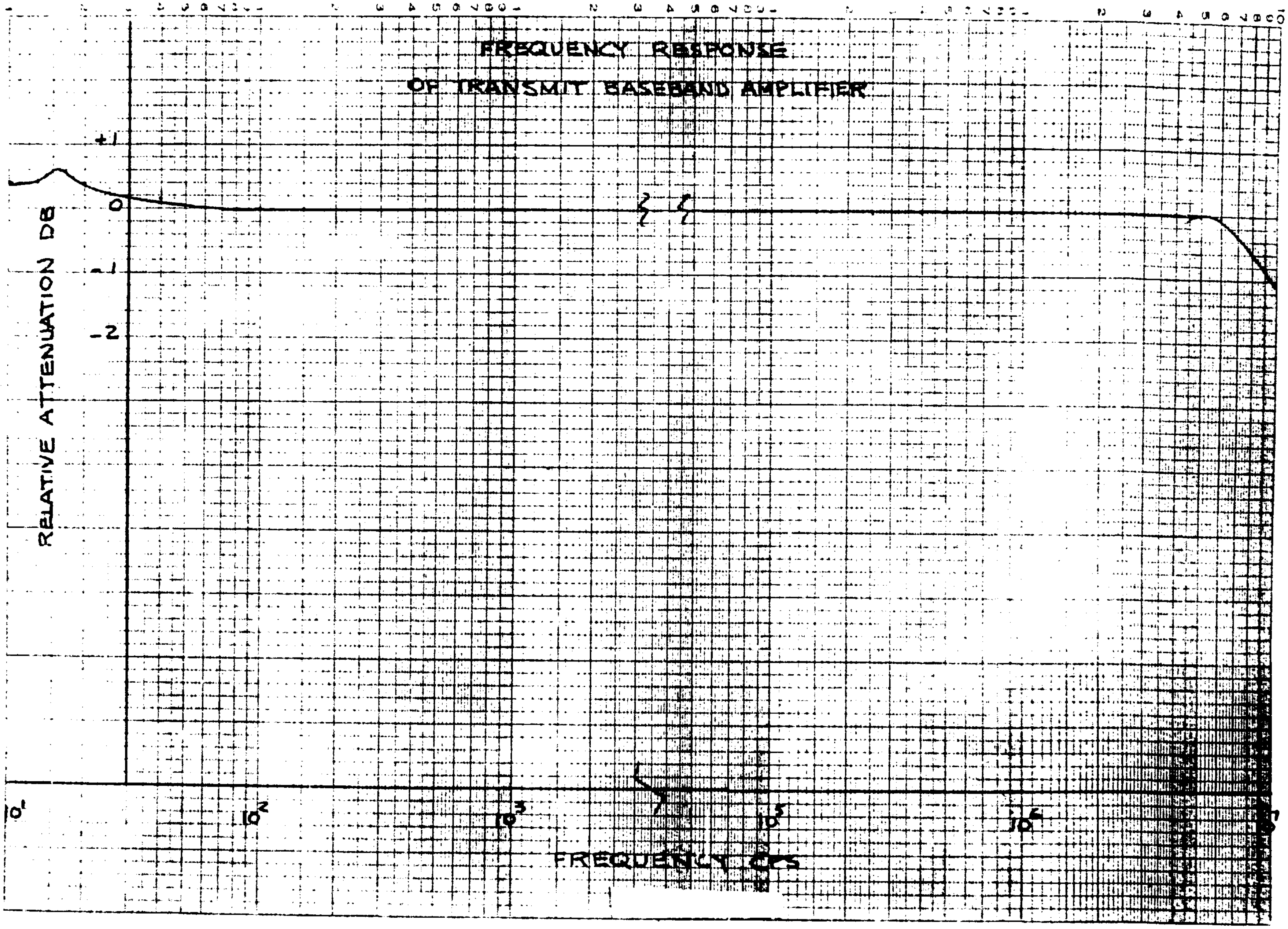
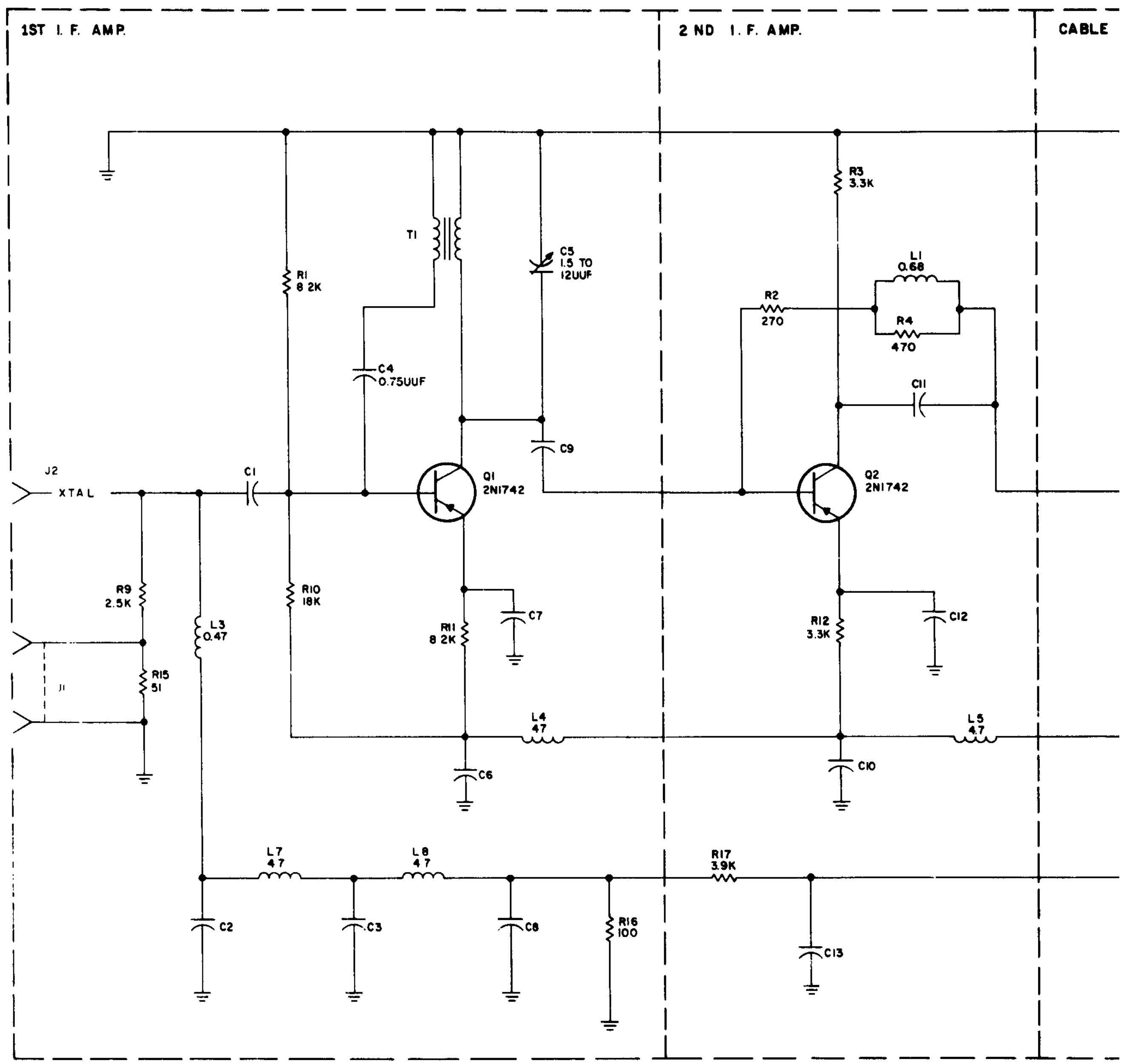
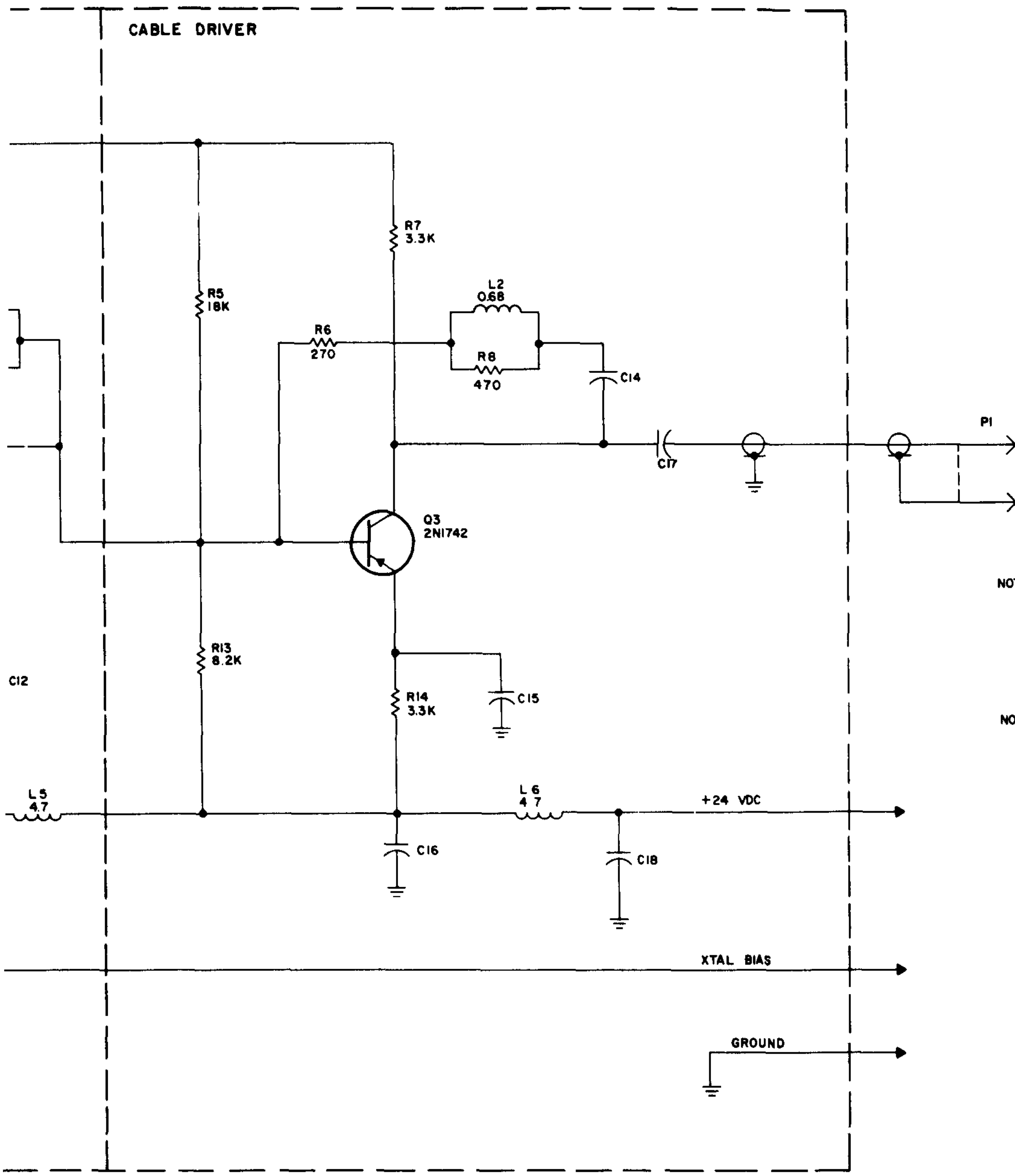


Figure 88 Frequency Response of Transmit Baseband Amplifier



1ST I. F. AMP.
2 ND I. F. AMP.
CABLE
T1
R1
8.2K
C4
0.75UUF
C5
1.5 TO
12UUF
C9
J2
XTAL
C1
Q1
2N1742
R9
2.5K
R10
18K
L3
0.47
R11
8.2K
C7
J1
R15
51
L4
47
C6
L7
47
L8
47
C2
C3
C8
R16
100
R3
3.3K
L1
0.68
R2
270
R4
470
C11
Q2
2N1742
R12
3.3K
C12
L5
4.7
C10
R17
3.9K
C13

1



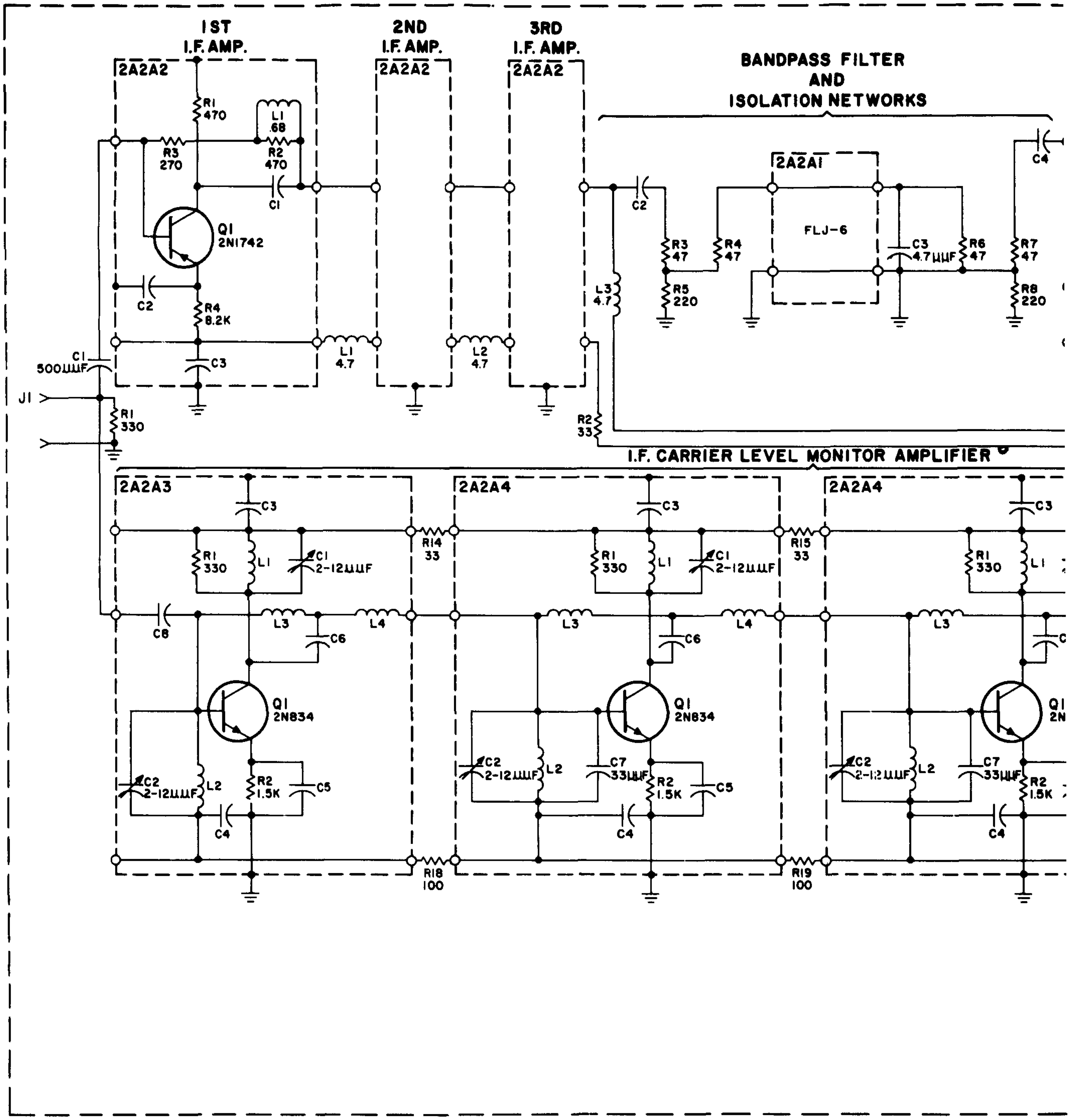
NOTE 1

UNLESS OTHERWISE INDICATED, ALL RESISTANCE VALUES ARE IN OHMS, ALL CAPACITOR VALUES ARE IN MICROFARADS AND ALL INDUCTANCE VALUES ARE IN MICROHENRIES UUF INDICATES MICROMICROFARADS AND K INDICATES THOUSANDS OF OHMS. ALL UNMARKED CAPACITORS ARE 0.001 MICROFARAD.

NOTE 2

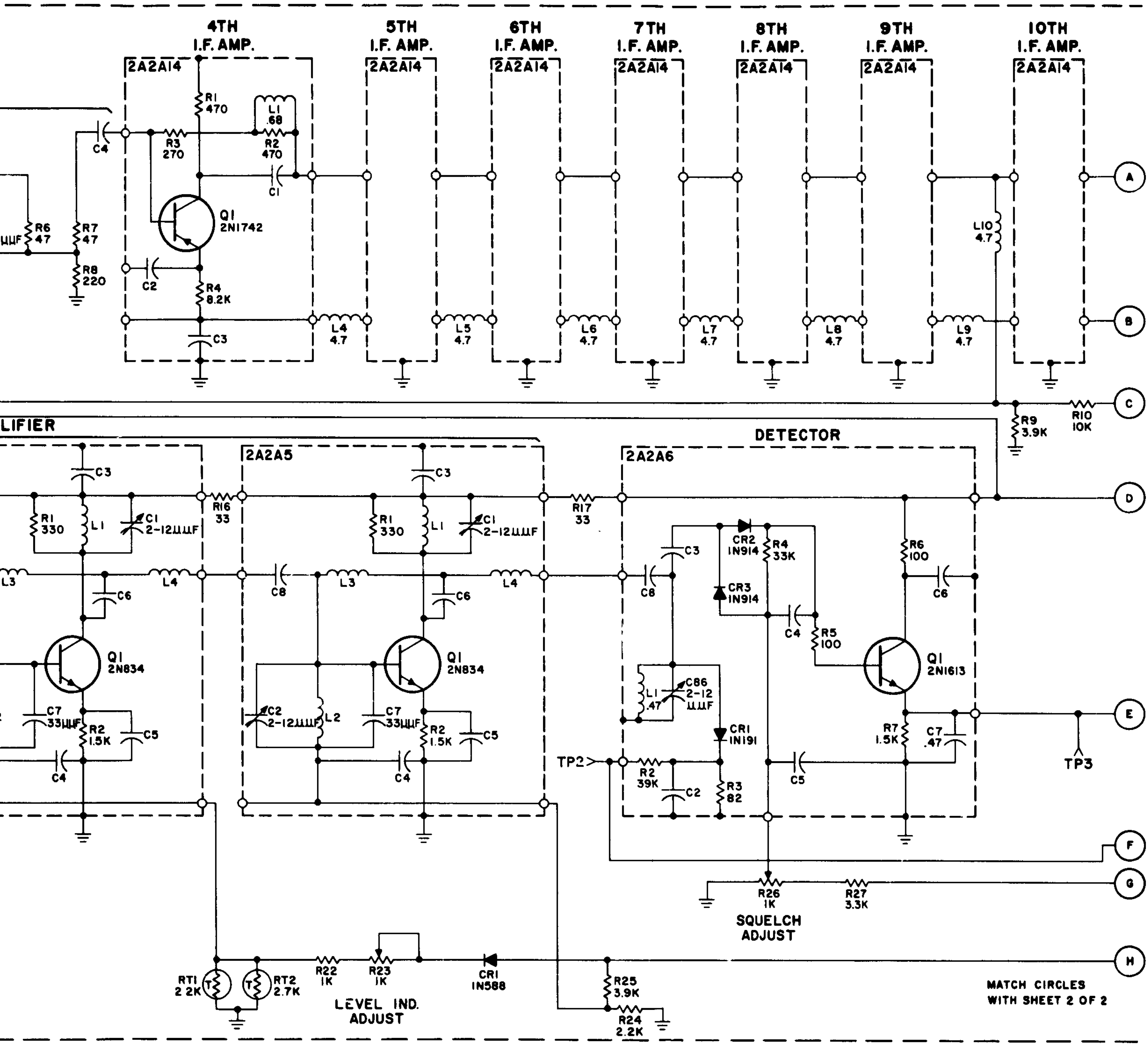
REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE DESIGNATIONS WITH 2A3

Figure 89 I-F Preamplifier, Schematic Diagram


1ST I.F. AMP.
2ND I.F. AMP.
3RD I.F. AMP.
2A2A2
BANDPASS FILTER AND ISOLATION NETWORKS
2A2A1
FLJ-6
I.F. CARRIER LEVEL MONITOR AMPLIFIER
2A2A3
2A2A4
Q1 2N1742
Q1 2N834
C1 500µµF
J1
R1 330
R2 33
R14 33
R15 33
R18 100
R19 100

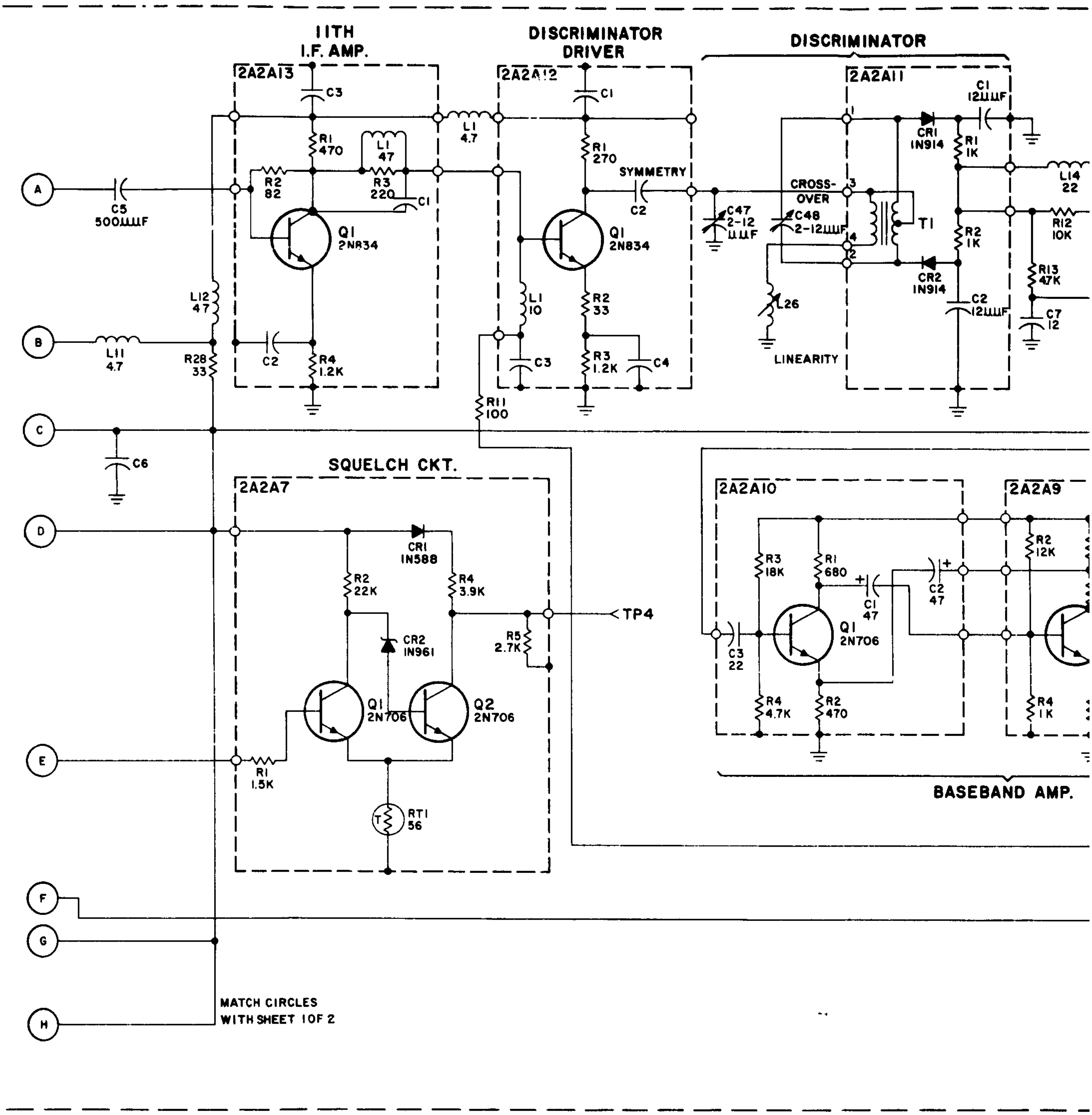
1

SHE



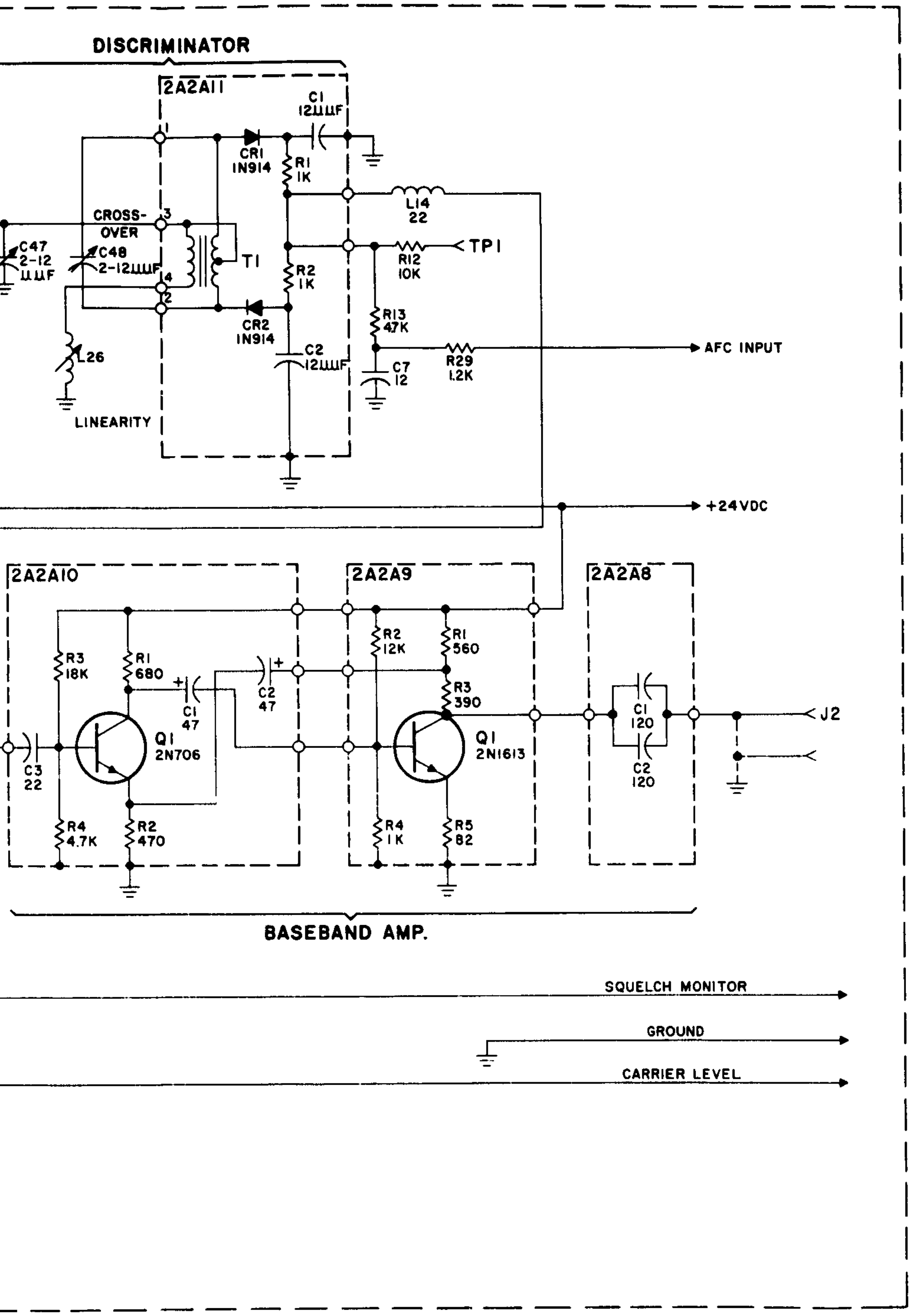
SHEET 1 OF 2

Figure 90 I-F Amplifier, Schematic Diagram


11TH I.F. AMP.
2A2A13
DISCRIMINATOR DRIVER
2A2A12
DISCRIMINATOR
2A2A11
SYMMETRY
CROSS-OVER
LINEARITY
SQUELCH CKT.
2A2A7
TP4
2A2A10
2A2A9
BASEBAND AMP.
MATCH CIRCLES WITH SHEET 1 OF 2

1

SHEET 2 OF 2



NOTE 1.

ALL RESISTOR VALUES ARE IN OHMS.
ALL INDUCTOR VALUES ARE IN MICROHENRIES.
ALL CAPACITOR VALUES ARE IN MICROFARADS UNLESS OTHERWISE INDICATED.
ALL UNMARKED CAPACITORS ARE .001 MICROFARAD.
μμF INDICATES MICROMICROFARAD.
K INDICATES THOUSANDS OF OHMS.
— — ○ — — CARD TERMINAL CONNECTION.
— — ● — — CARD GROUND CONNECTION.

NOTE 2.

ALL REFERENCE DESIGNATIONS ARE ABBREVIATED.
DESIGNATIONS APPEARING WITHIN DOTTED OUTLINES SHOULD BE PREFIXED WITH THE NUMBER DIRECTLY INSIDE CORRESPONDING OUTLINE.
ALL OTHER DESIGNATIONS SHOULD BE PREFIXED WITH 2A2

SHEET 2 OF 2

Figure 90 I-F Amplifier, Schematic Diagram

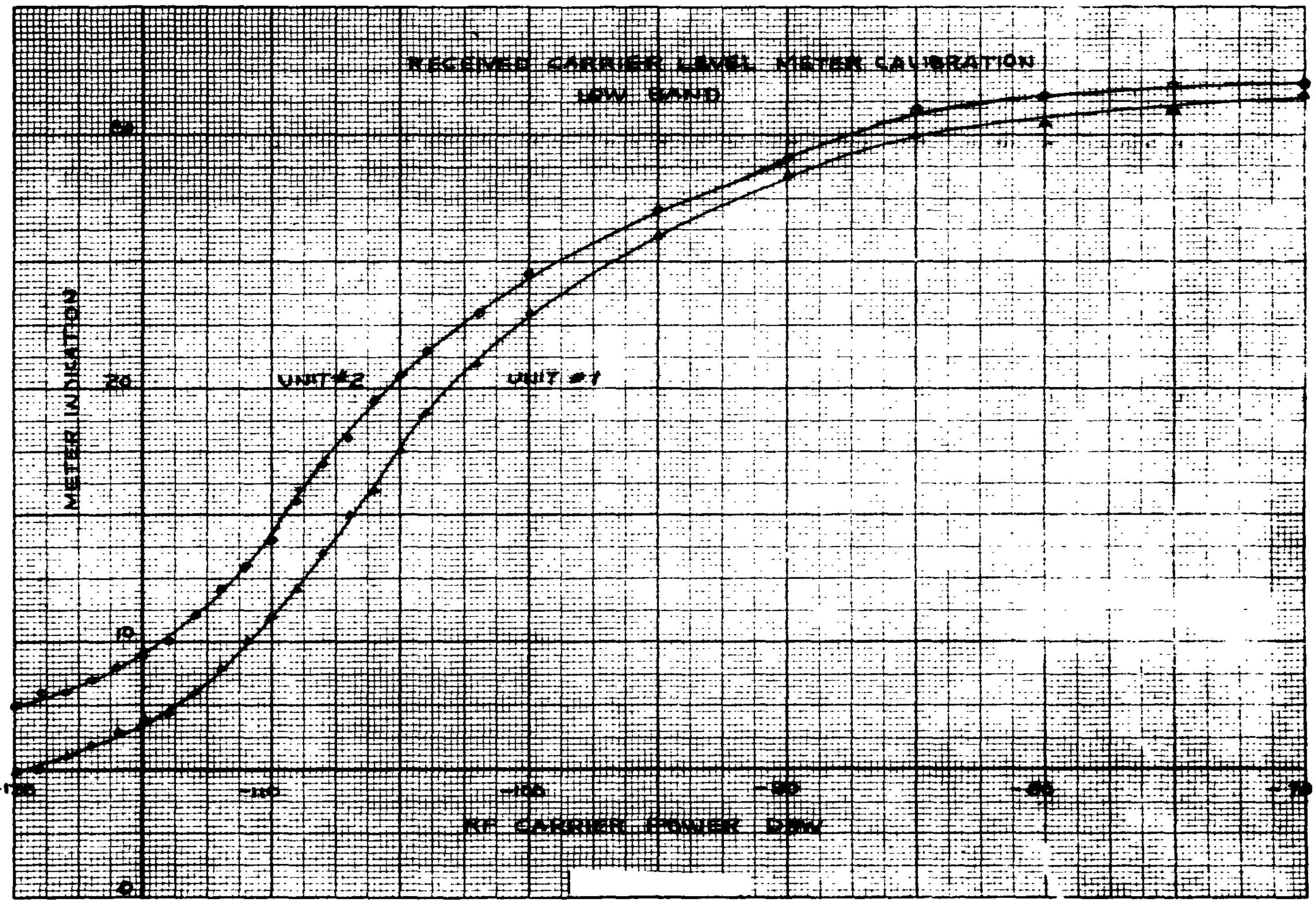


Figure 91 Received Carrier Level Meter Calibration, Low Band

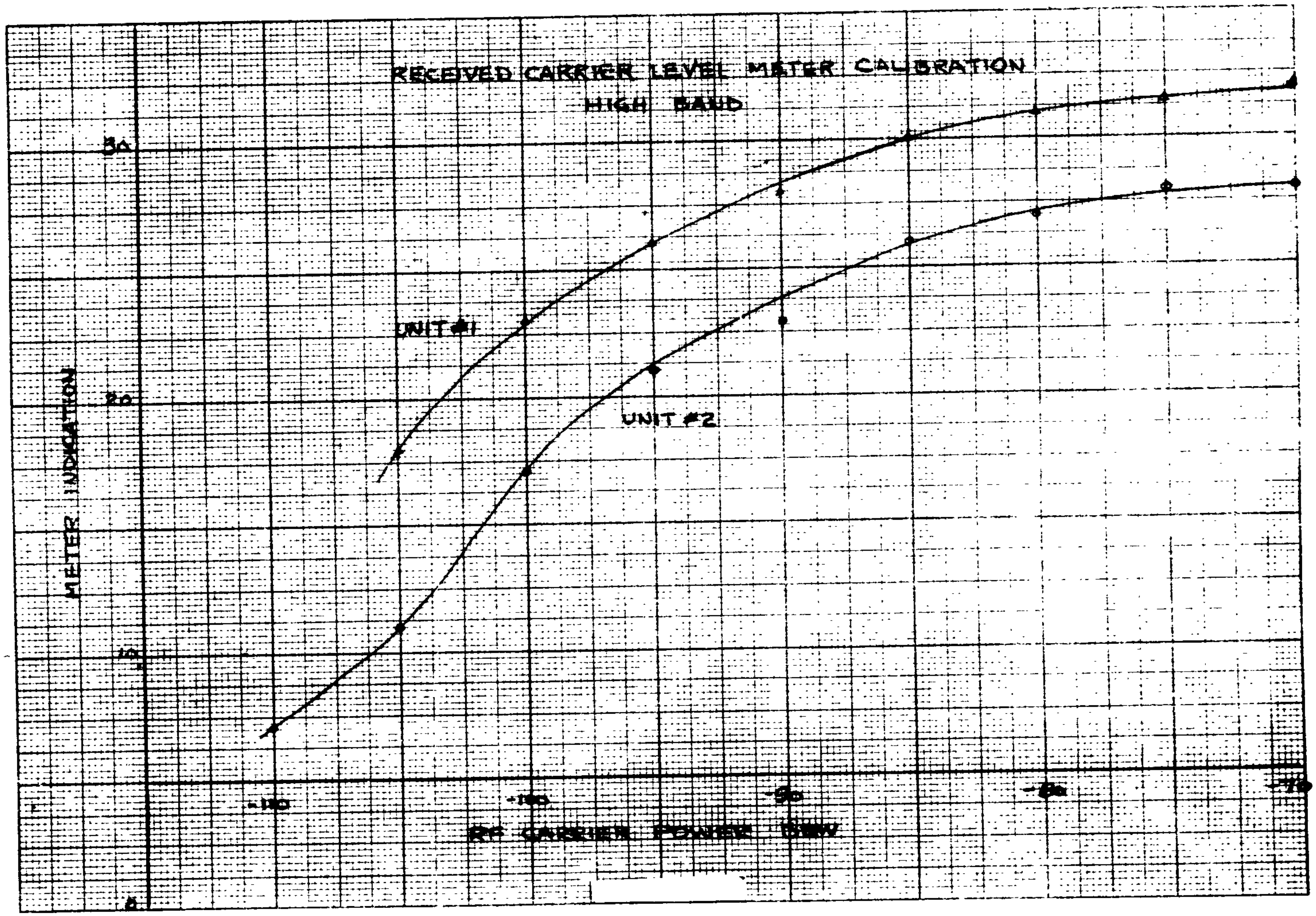
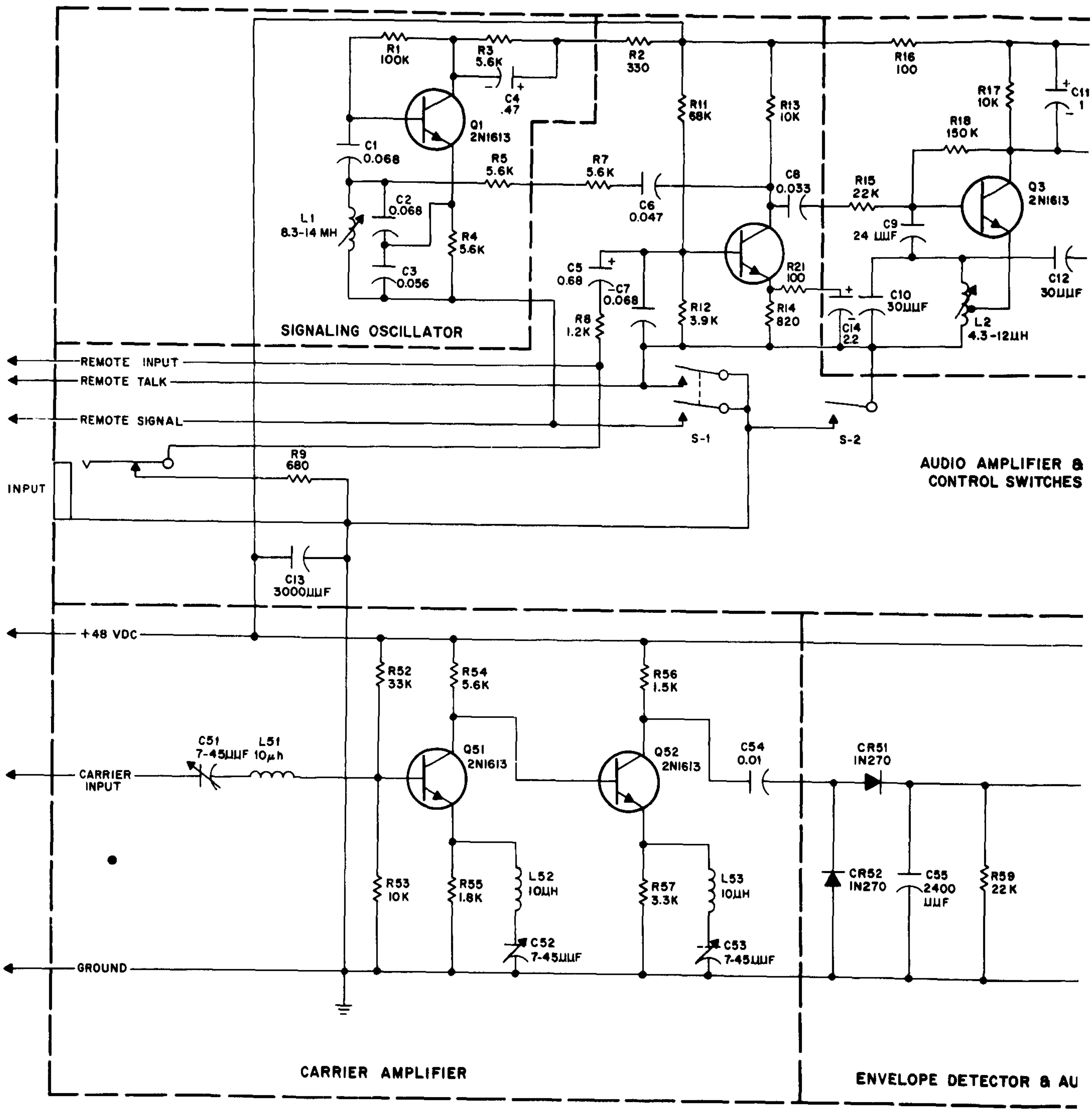
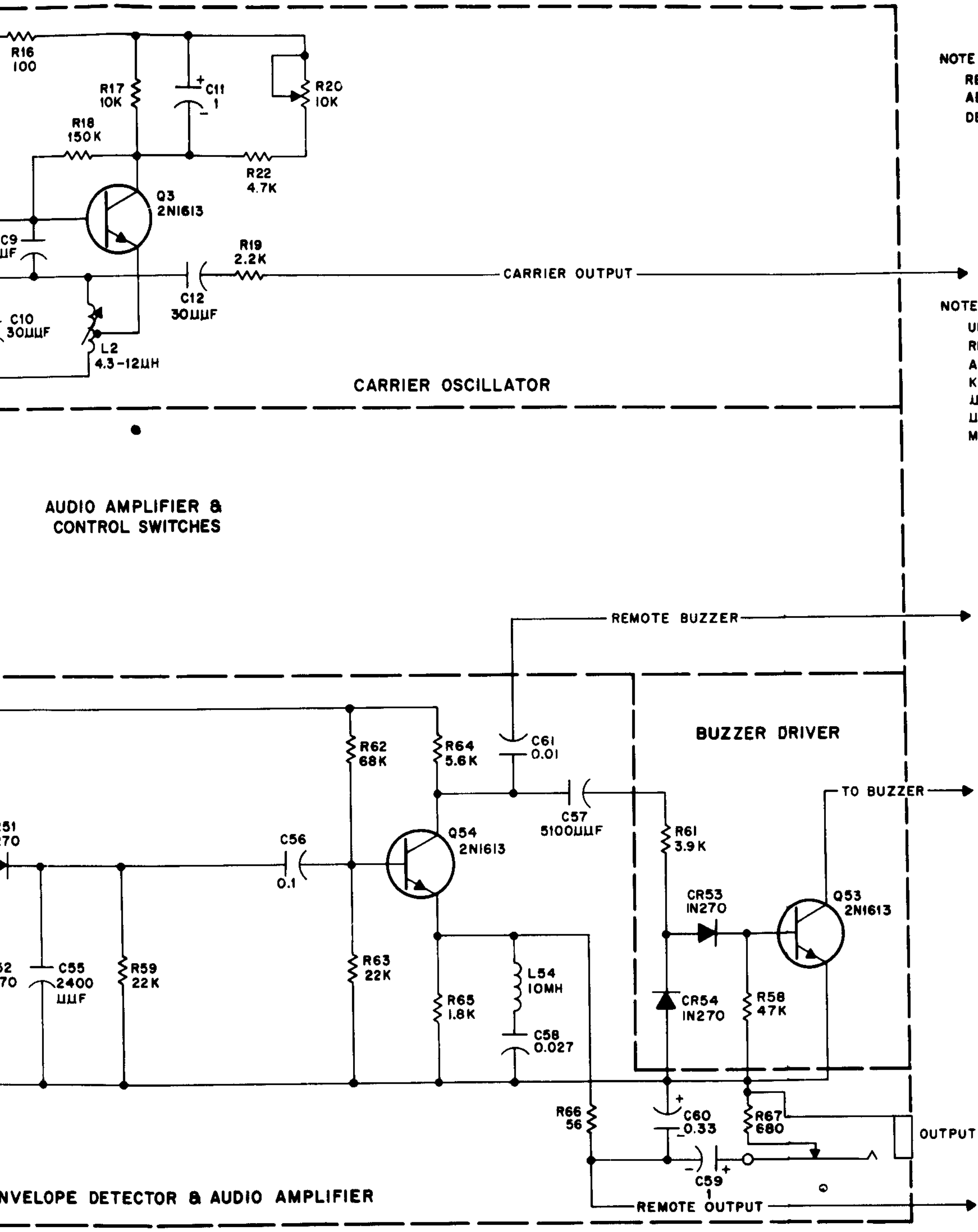


Figure 92 Received Carrier Level Meter Calibration, High Band


SIGNALING OSCILLATOR
AUDIO AMPLIFIER &
CONTROL SWITCHES
CARRIER AMPLIFIER
ENVELOPE DETECTOR & AU
R1 100K
R3 5.6K
R2 330
R16 100
R17 10K
R18 150K
C11 1
C4 .47
Q1 2N1613
R11 68K
R13 10K
C1 0.068
R5 5.6K
R7 5.6K
C8 0.033
R15 22K
Q3 2N1613
C2 0.068
C6 0.047
L1 8.3-14 MH
R4 5.6K
C9 24 UUF
C3 0.056
C5 0.68
R21 100
C12 30UUF
C7 0.068
C10 30UUF
R8 1.2K
R12 3.9K
R14 820
C14 22
L2 4.3-12UH
REMOTE INPUT
REMOTE TALK
REMOTE SIGNAL
S-1
S-2
R9 680
INPUT
C13 3000UUF
+48 VDC
R52 33K
R54 5.6K
R56 1.5K
C51 7-45UUF
L51 10μh
Q51 2N1613
Q52 2N1613
C54 0.01
CR51 1N270
CARRIER INPUT
R53 10K
R55 1.8K
L52 10UH
R57 3.3K
L53 10UH
CR52 1N270
C55 2400 UUF
R59 22K
C52 7-45UUF
C53 7-45UUF
GROUND

1



NOTE 1

REFERENCE DESIGNATIONS ARE ABBREVIATED PREFIX THE DESIGNATIONS WITH 2A7

NOTE 2

UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS
ALL CAPACITANCE VALUES ARE IN MICROFARADS
K INDICATES THOUSANDS OF OHMS
UUF INDICATES MICROMICROFARADS
UH INDICATES MICROHENRIES
MH INDICATES MILLIHENRIES

Figure 93 Order Wire, Schematic Diagram

NOTE 1
REFERENCE DESIGNATIONS
ARE ABBREVIATED. PREFIX
THE DESIGNATIONS WITH 2A5

NOTE 2
UNLESS OTHERWISE INDICATED
ALL RESISTANCE VALUES ARE IN OHMS
ALL CAPACITANCE VALUES
ARE IN MICROFARADS
K INDICATES THOUSAND OF OHMS
µµF INDICATES MICROMICROFARADS
UH INDICATES MICROHENRIES

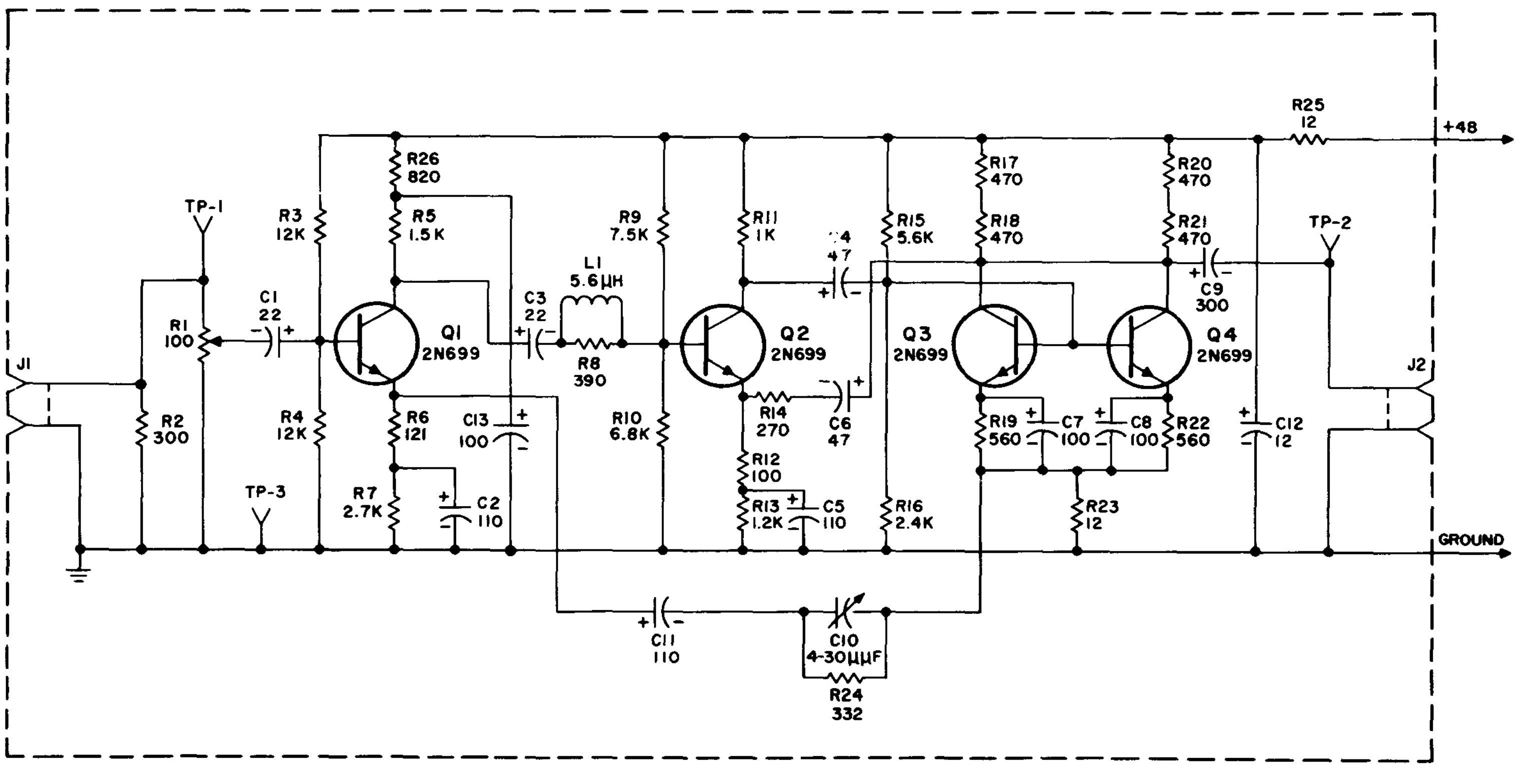
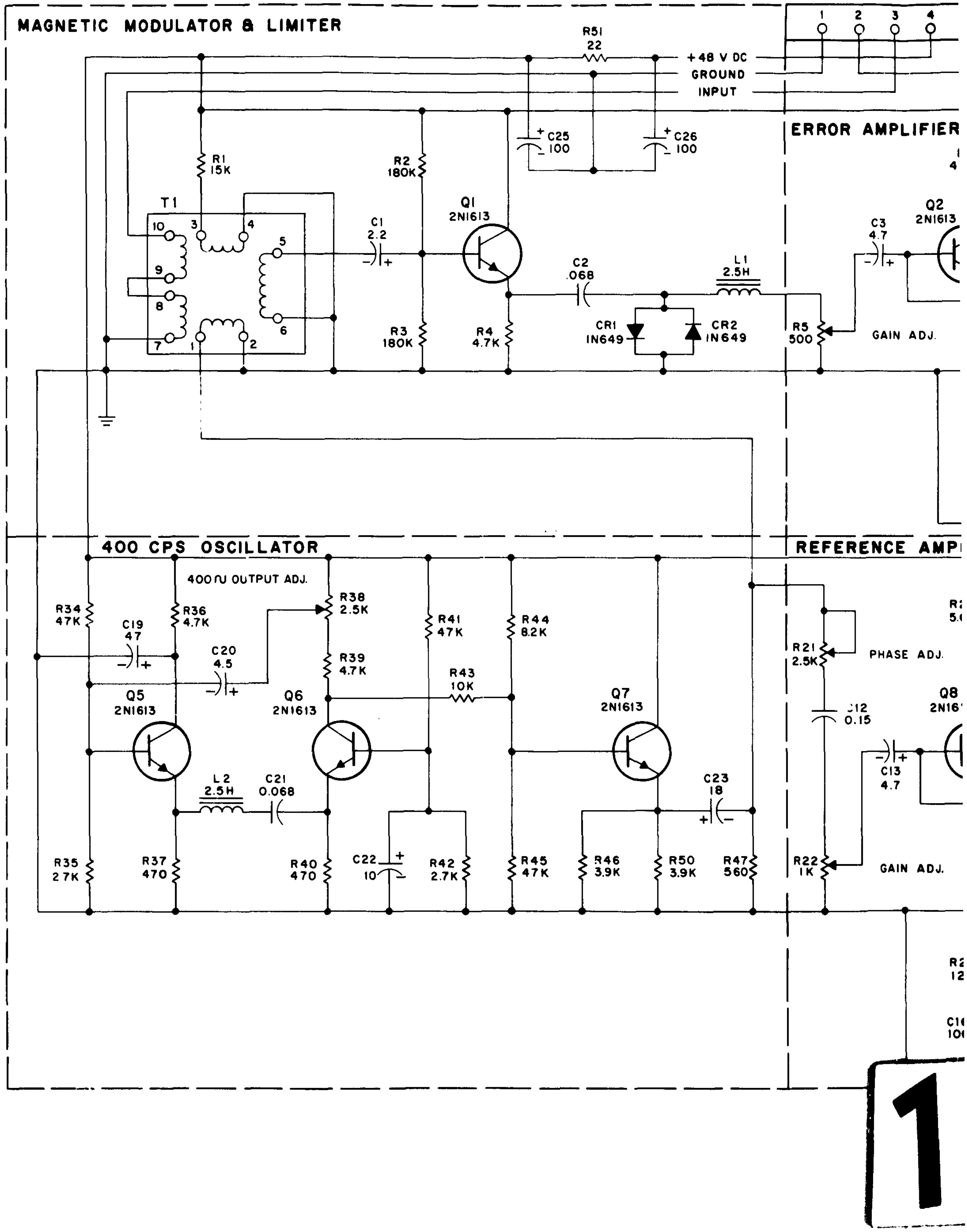


Figure 94 Receive Baseband Amplifier, Schematic Diagram


MAGNETIC MODULATOR & LIMITER
R51 22
+48 V DC
GROUND
INPUT
1 2 3 4
ERROR AMPLIFIER
C25 100
C26 100
R1 15K
R2 180K
T1
Q1 2N1613
C1 2.2
C2 .068
L1 2.5H
CR1 1N649
CR2 1N649
R3 180K
R4 4.7K
R5 500
GAIN ADJ.
C3 4.7
Q2 2N1613
400 CPS OSCILLATOR
400 ∿ OUTPUT ADJ.
R34 47K
R36 4.7K
C19 47
C20 4.5
R38 2.5K
R39 4.7K
R41 47K
R44 8.2K
R43 10K
Q5 2N1613
Q6 2N1613
Q7 2N1613
L2 2.5H
C21 0.068
C23 18
R35 2.7K
R37 470
R40 470
C22 10
R42 2.7K
R45 47K
R46 3.9K
R50 3.9K
R47 560
REFERENCE AMP
R21 2.5K
PHASE ADJ.
C12 0.15
C13 4.7
Q8
R22 1K
GAIN ADJ.
1

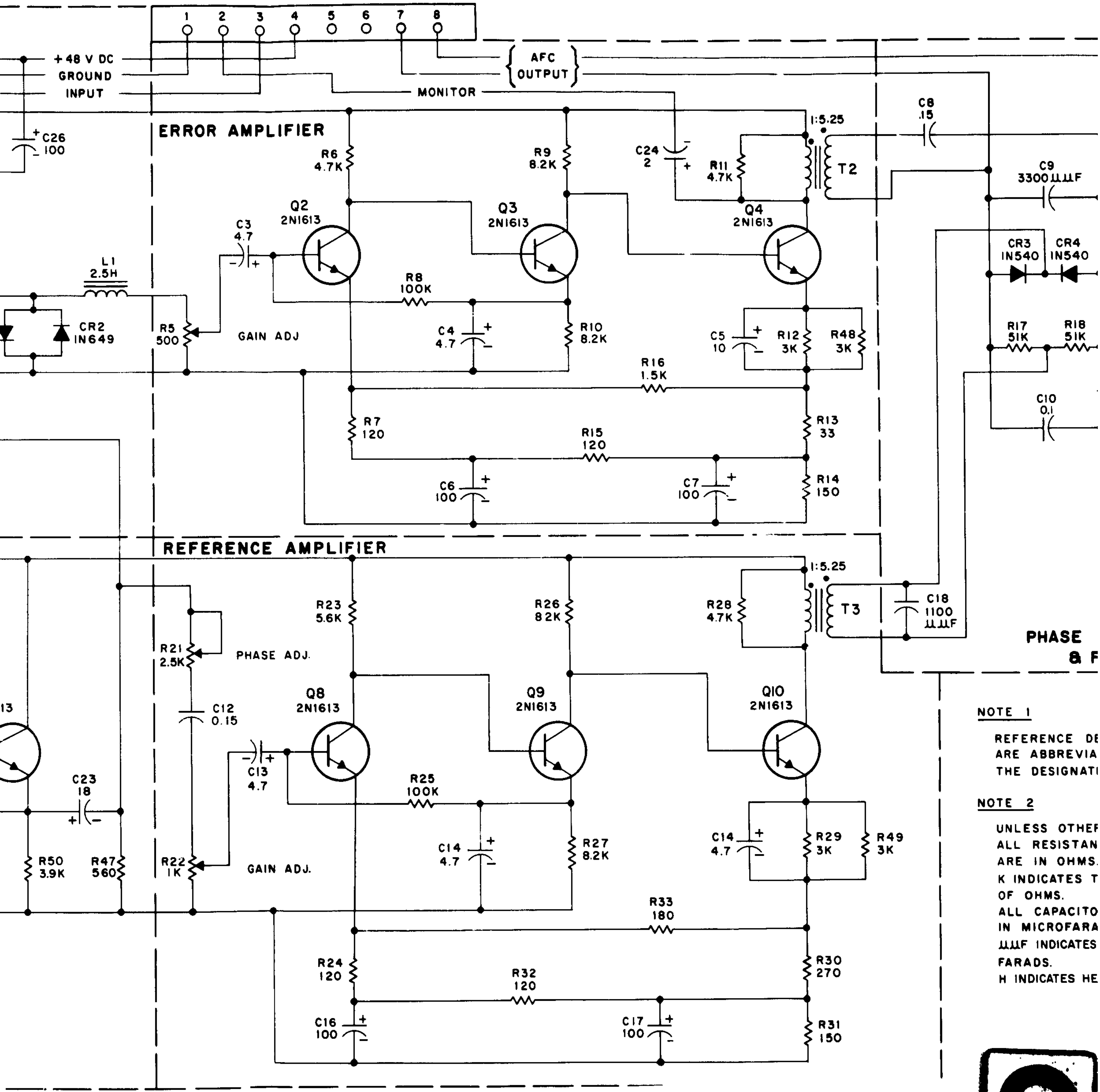
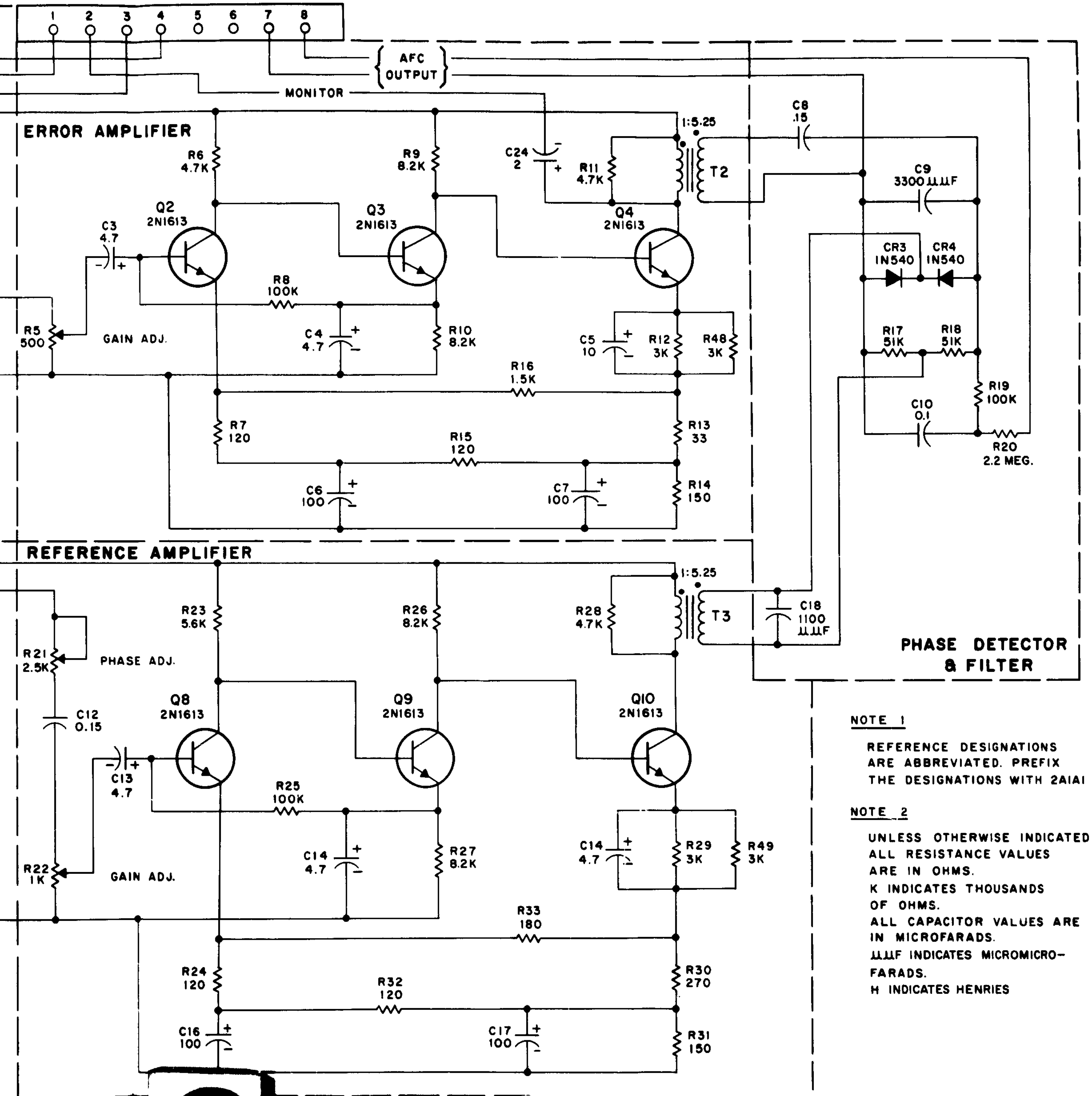


Figure 95 Automatic Frequency Cont
Diagram

2



3

Figure 95 Automatic Frequency Control Amplifier, Schematic Diagram

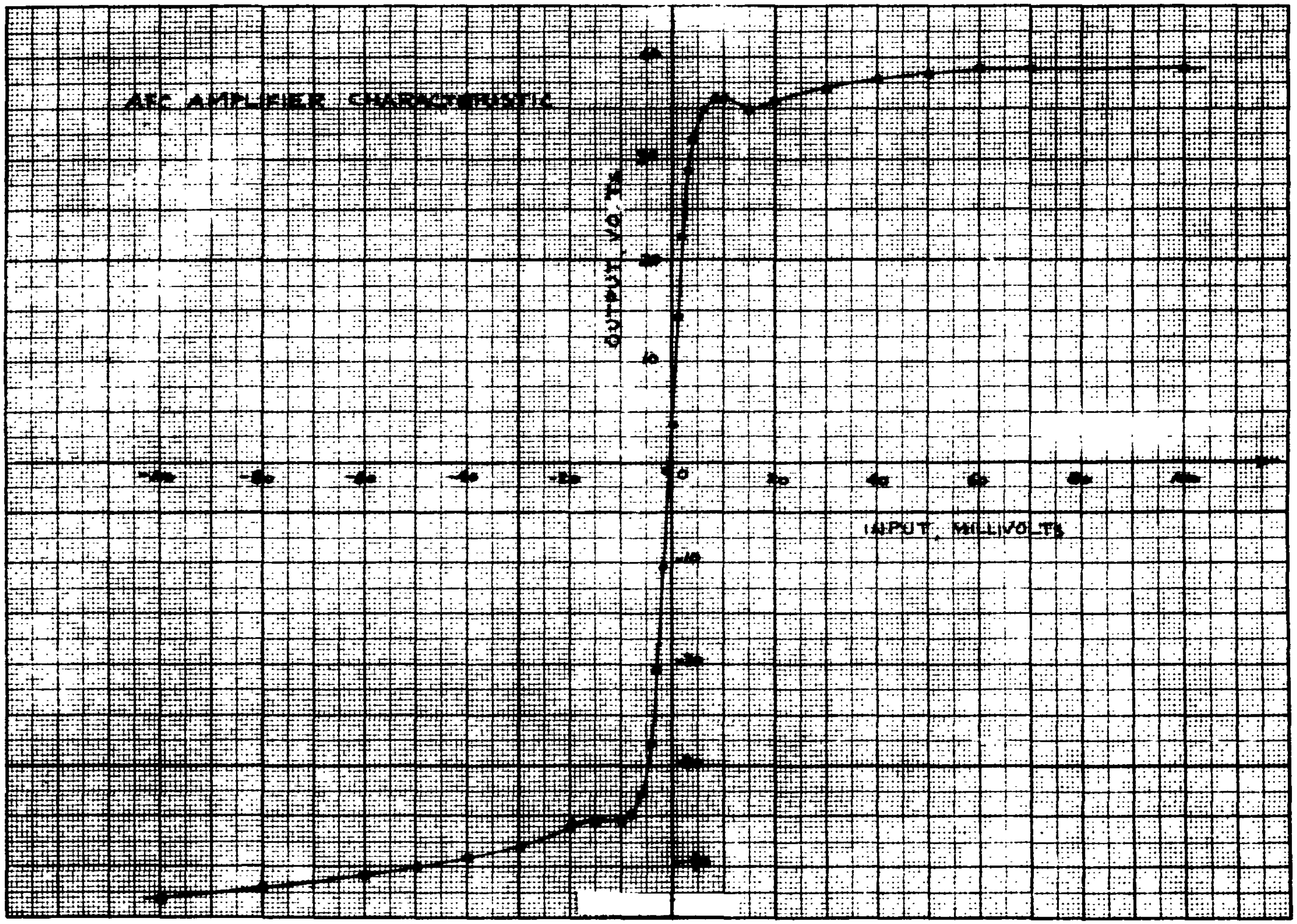
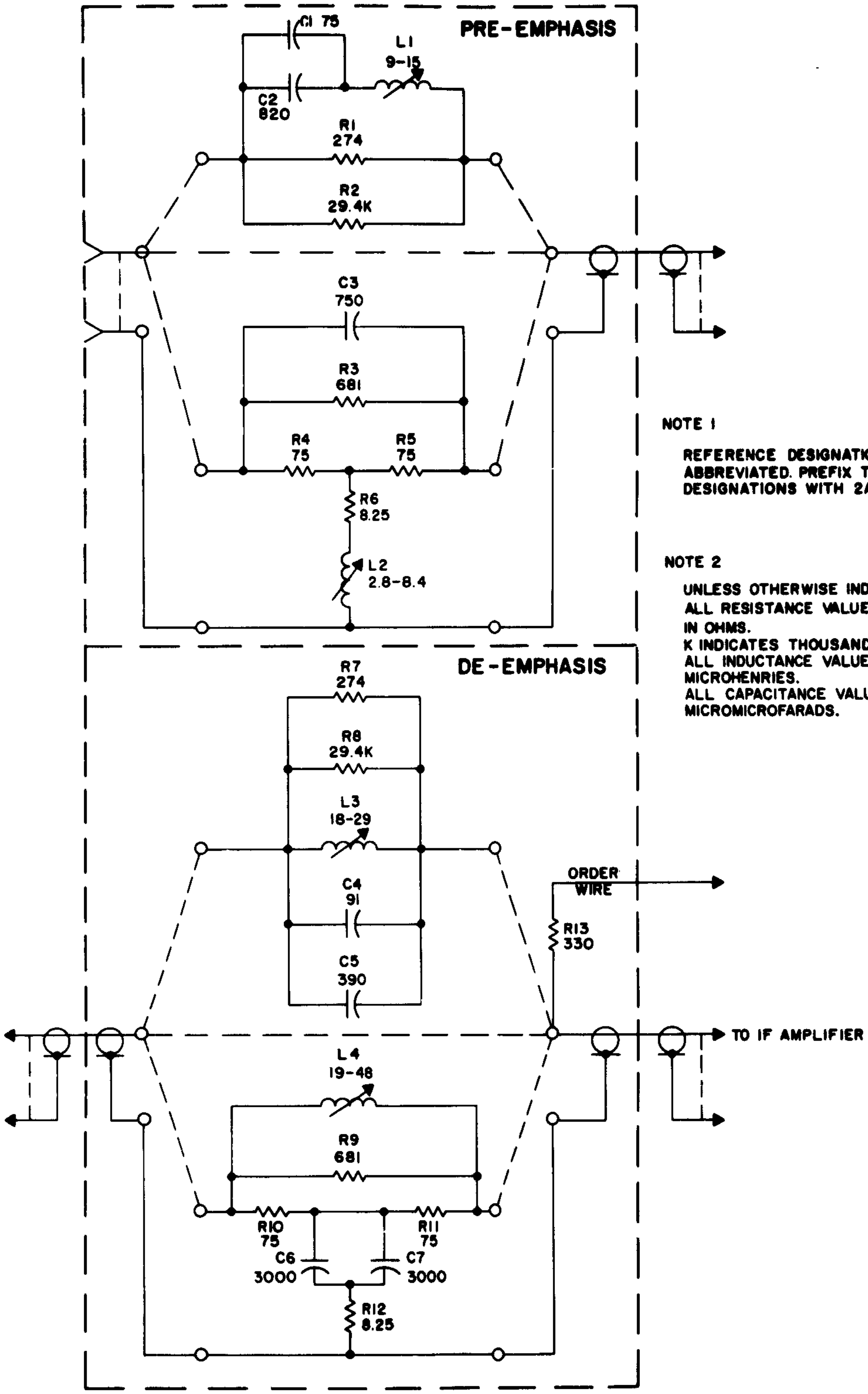


Figure 96 AFC Amplifier Transfer Characteristic



NOTE 1

REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE DESIGNATIONS WITH 2A6

NOTE 2

UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS.
K INDICATES THOUSANDS OF OHMS.
ALL INDUCTANCE VALUES ARE IN MICROHENRIES.
ALL CAPACITANCE VALUES ARE IN MICROMICROFARADS.

Figure 97 Pre- and De-emphasis Circuit, Schematic Diagram

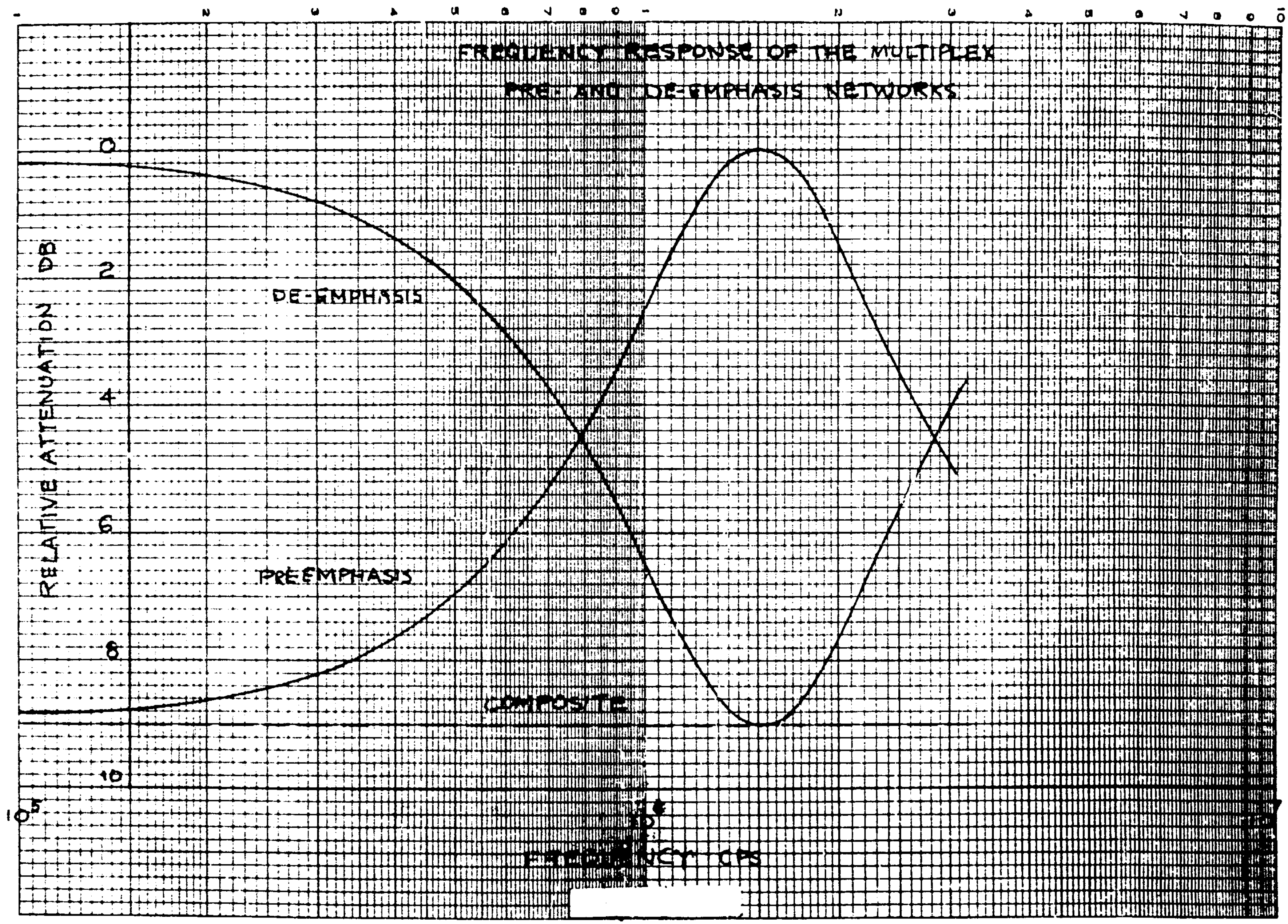


Figure 98 Frequency Response of Multiplex Pre- and De-emphasis Networks

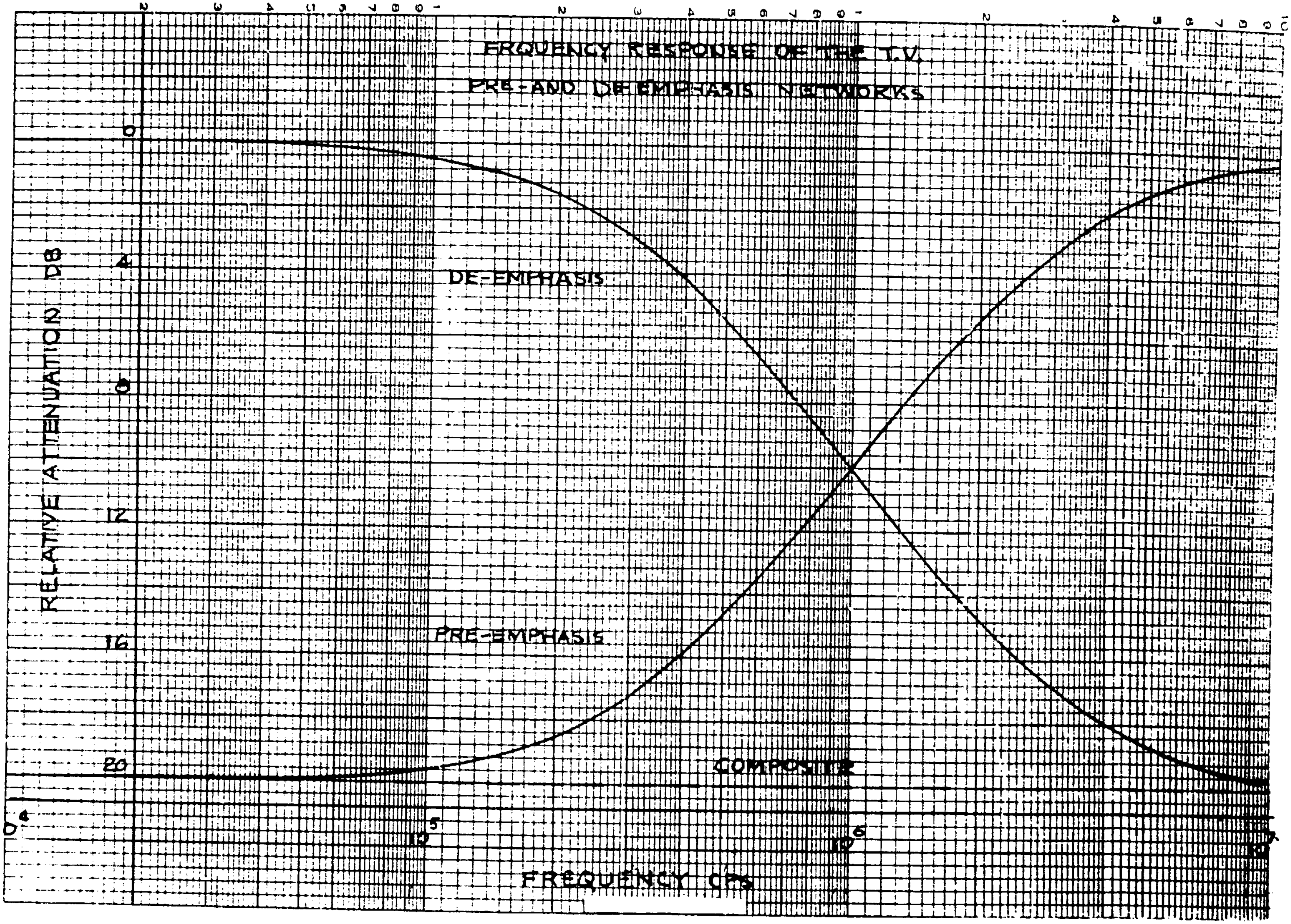
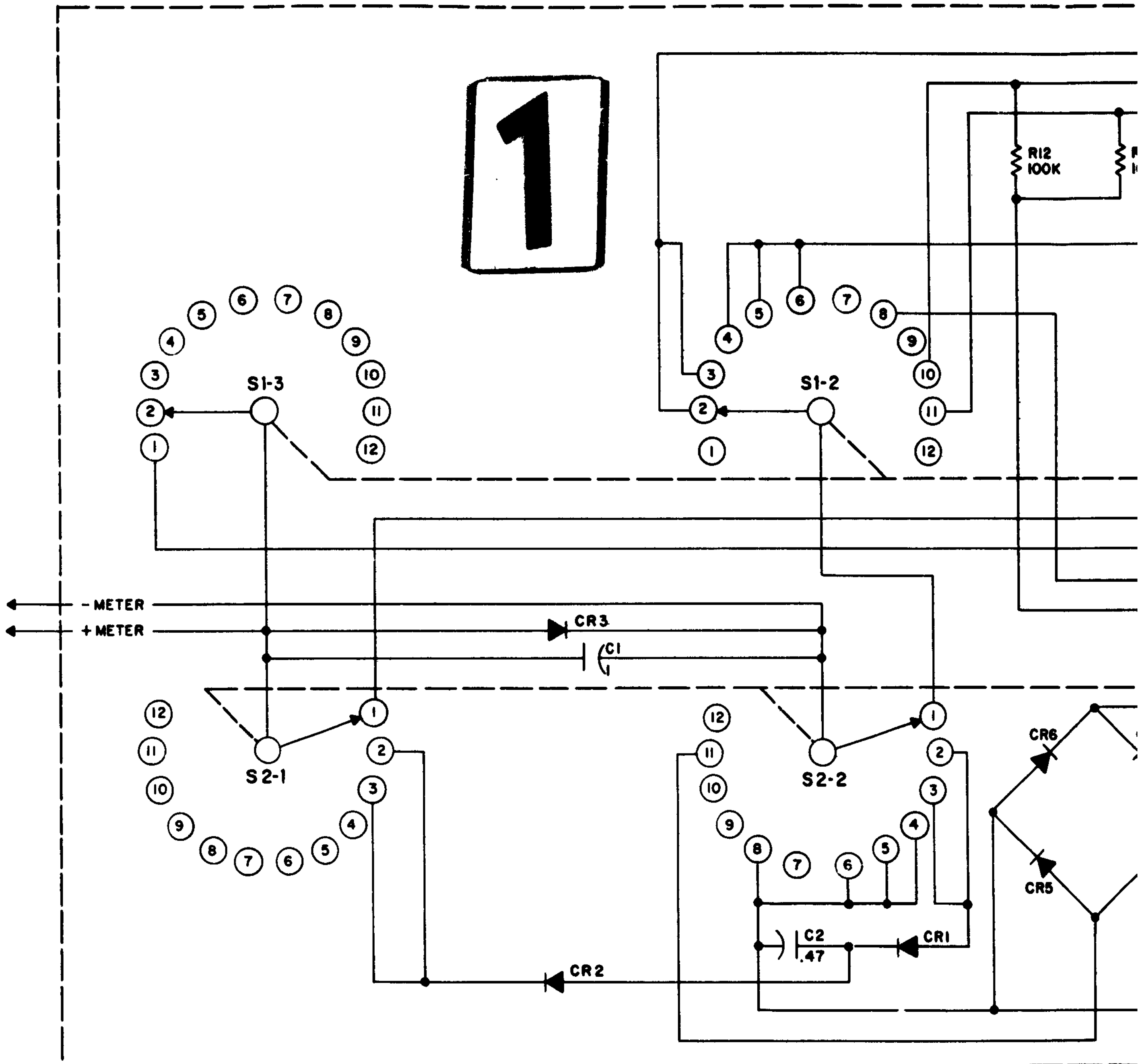


Figure 99 Frequency Response of TV Pre- and De-emphasis Networks



POSITION S1	POSITION S2	FUNCTION
1	1	OFF
2	1	FILAMENT VOLTAGE RCVR KLYS
3	1	FILAMENT VOLTAGE XMTR KLYS
4	1	+24 V
5	1	+48 V
6	1	+90 V
7	1	OFF
8	1	KLYS CATHODE VOLTAGE
9	1	OFF
10	1	REFLECTOR VOLTAGE RCVR KLYS
11	1	REFLECTOR VOLTAGE XMTR KLYS
12	1	OFF

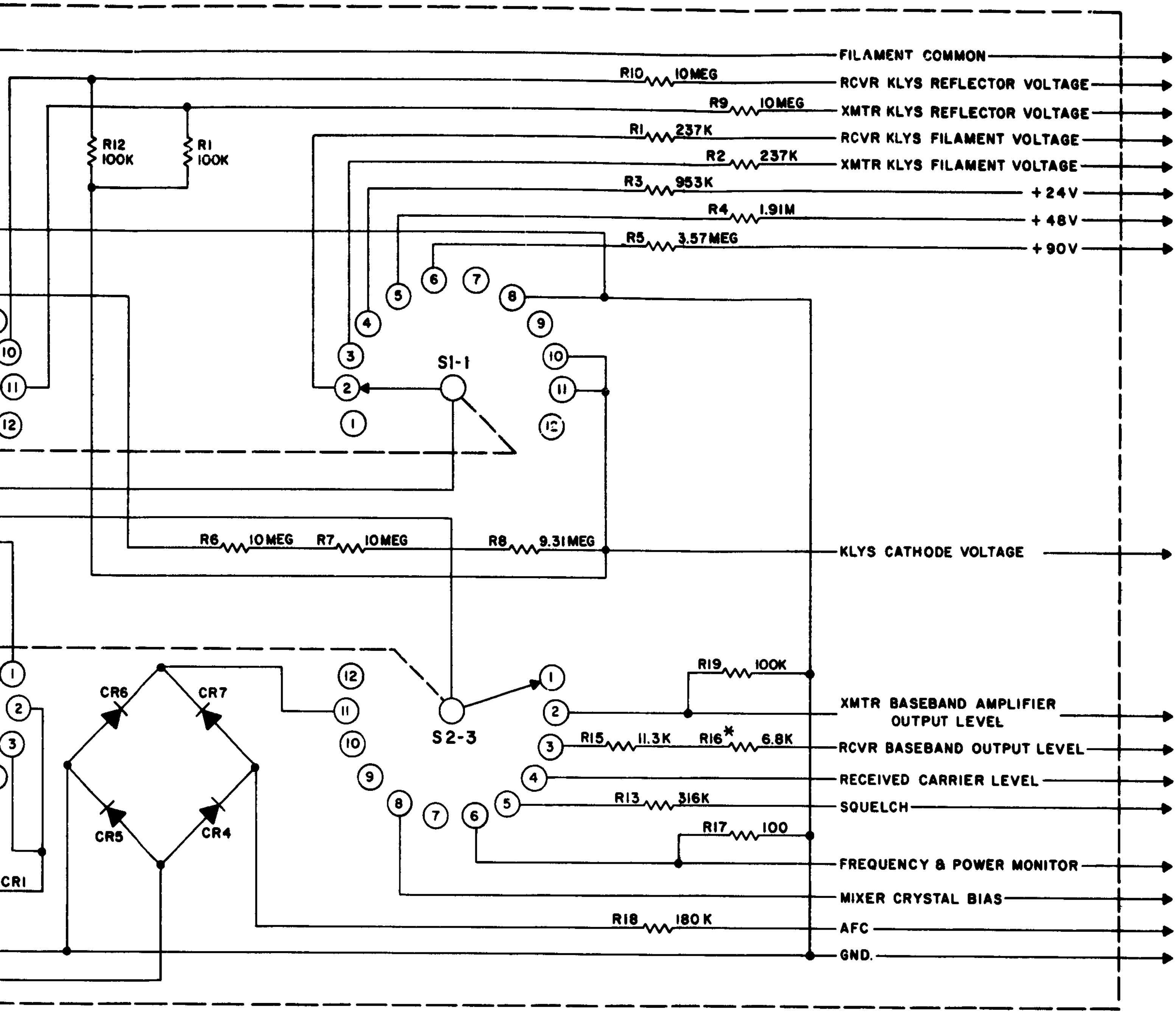
POSITION S1	POSITION S2	FUNCTION
1	1	OFF
1	2	XMTR DEVIATION
1	3	RCVR BASEBAND LEVEL
1	4	RCVD CARRIER LEVEL
1	5	SQUELCH
1	6	XMTR FREQ. & POWER
1	7	OFF
1	8	CRYSTAL BIAS
1	9	OFF
1	10	OFF
1	11	AFC
1	12	OFF

NOTE 1

REFERENCE DESIGNATION
PREFIX THE DESIGNATIO

NOTE 2

UNLESS OTHERWISE IND
VALUES ARE IN OHMS
ALL CAPACITANCE VALU
K INDICATES THOUSAND
MEG INDICATES MILLIONS



E1

REFERENCE DESIGNATIONS ARE ABBREVIATED. PREFIX THE DESIGNATIONS WITH 2A8

E2

UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS

ALL CAPACITANCE VALUES ARE IN MICROFARADS

K INDICATES THOUSANDS OF OHMS

MEG INDICATES MILLIONS OF OHMS

Figure 100 Metering and Test Facility, Schematic Diagram

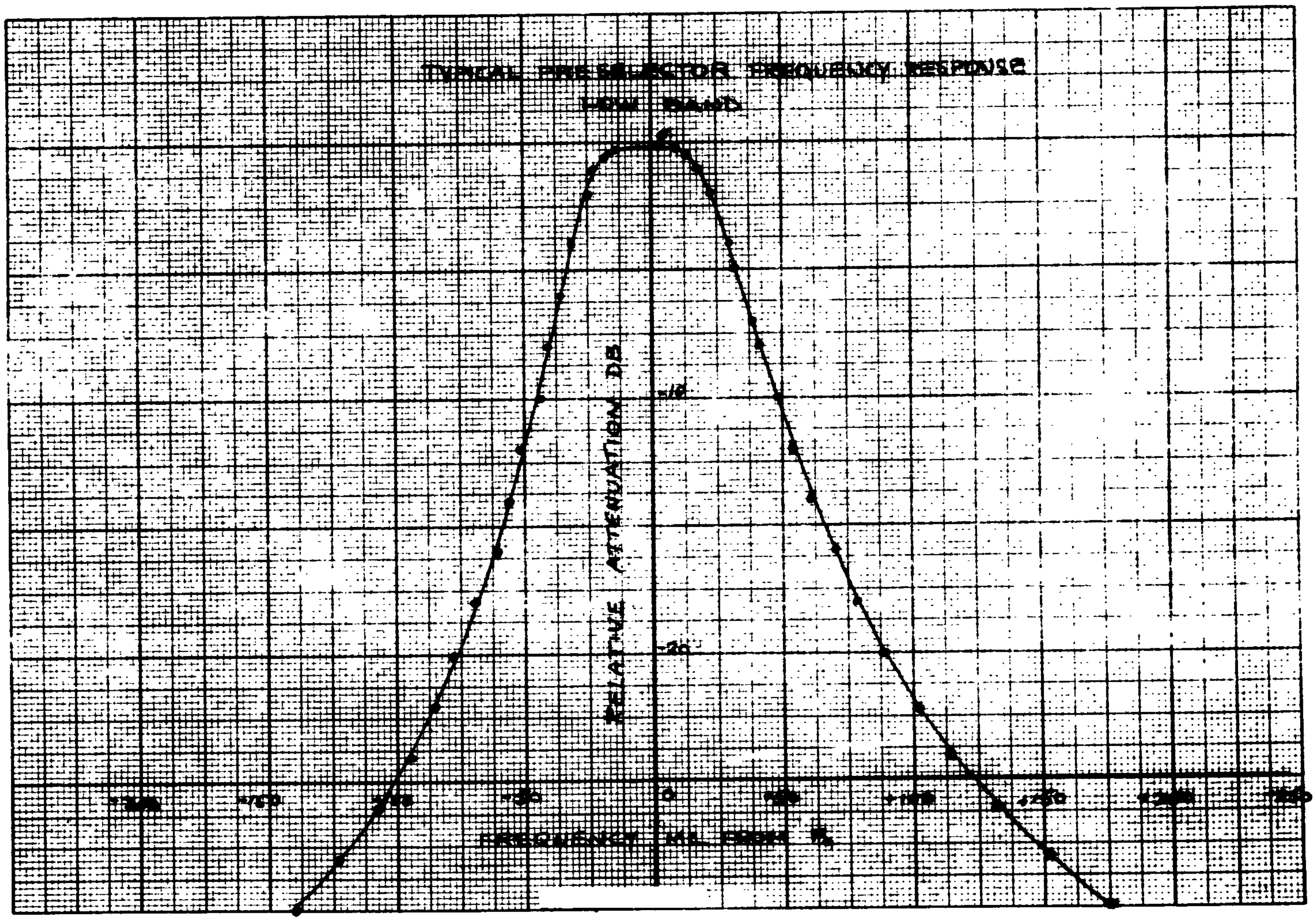


Figure 101 Typical Preselector Frequency Response, Low Band

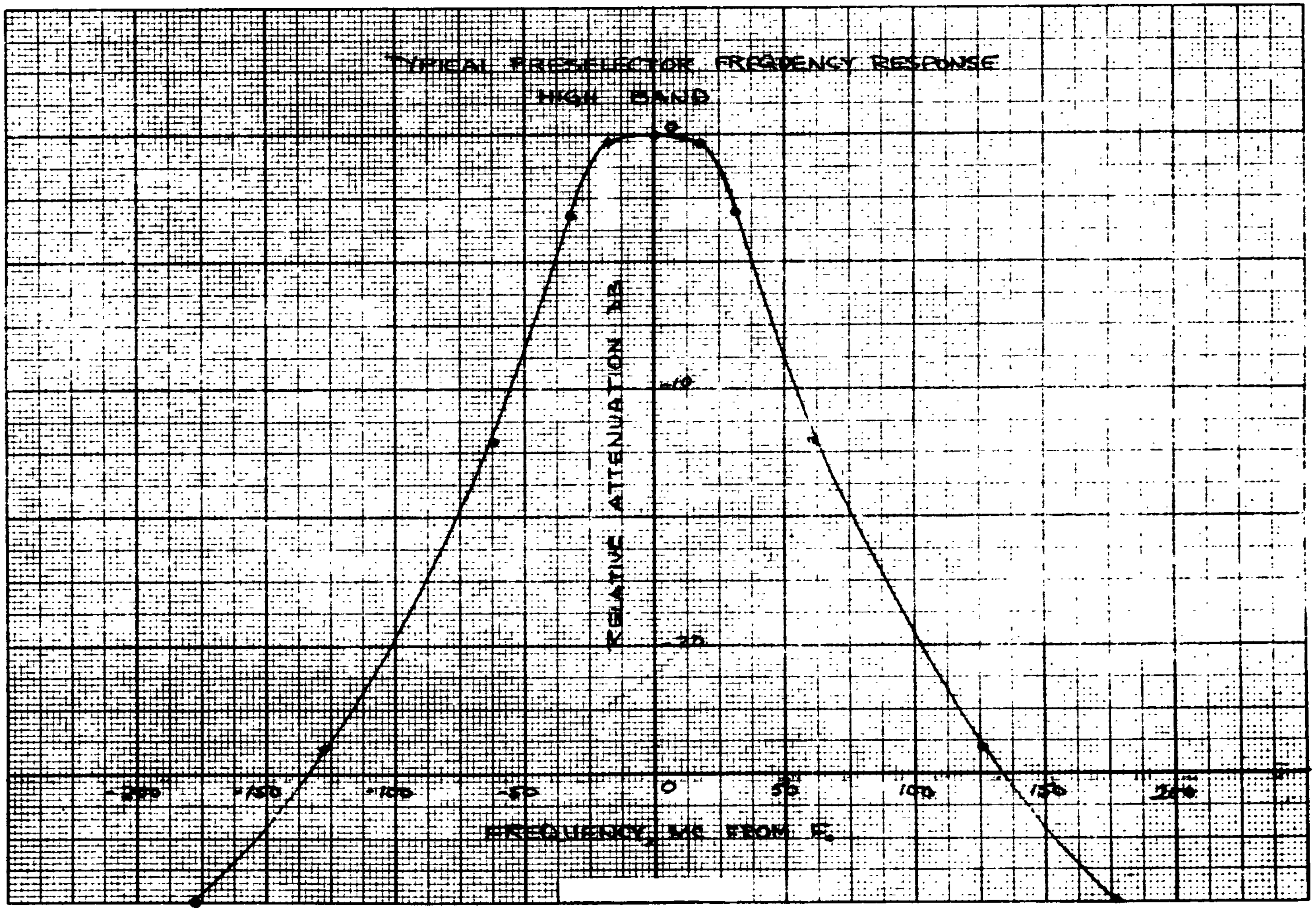


Figure 102 Typical Preselector Frequency Response, High Band

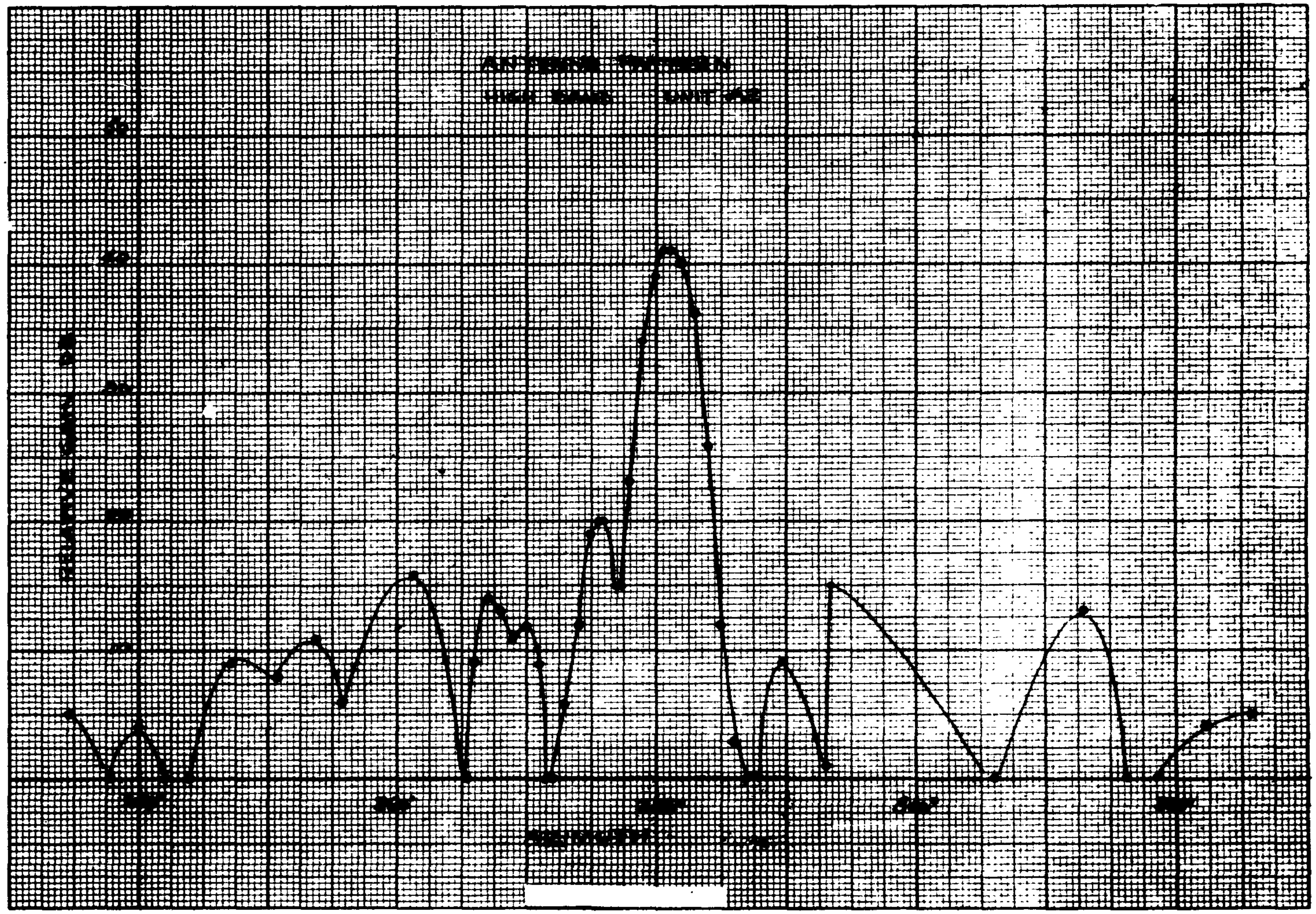


Figure 103 Antenna Pattern, High Band, Unit No. 2

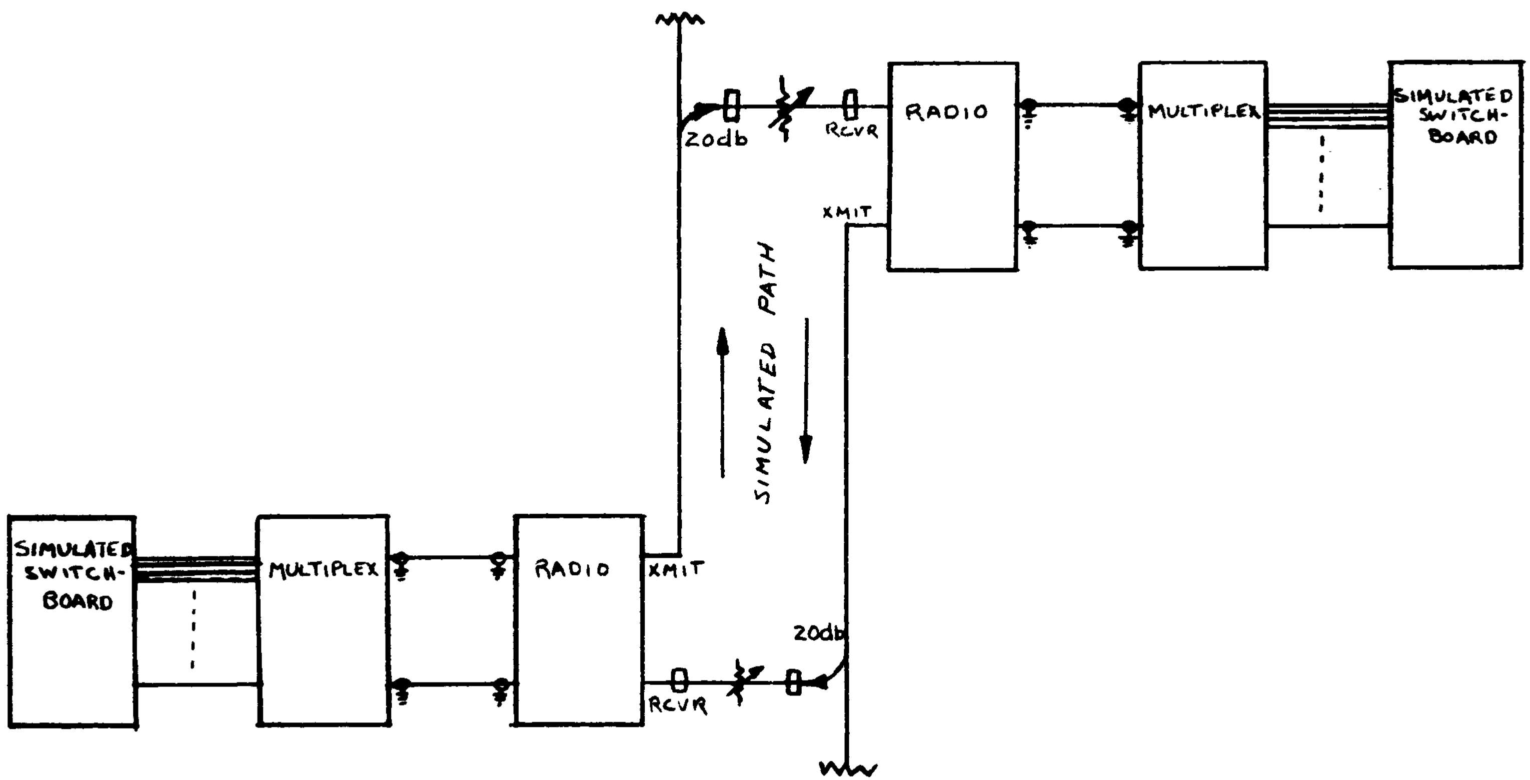
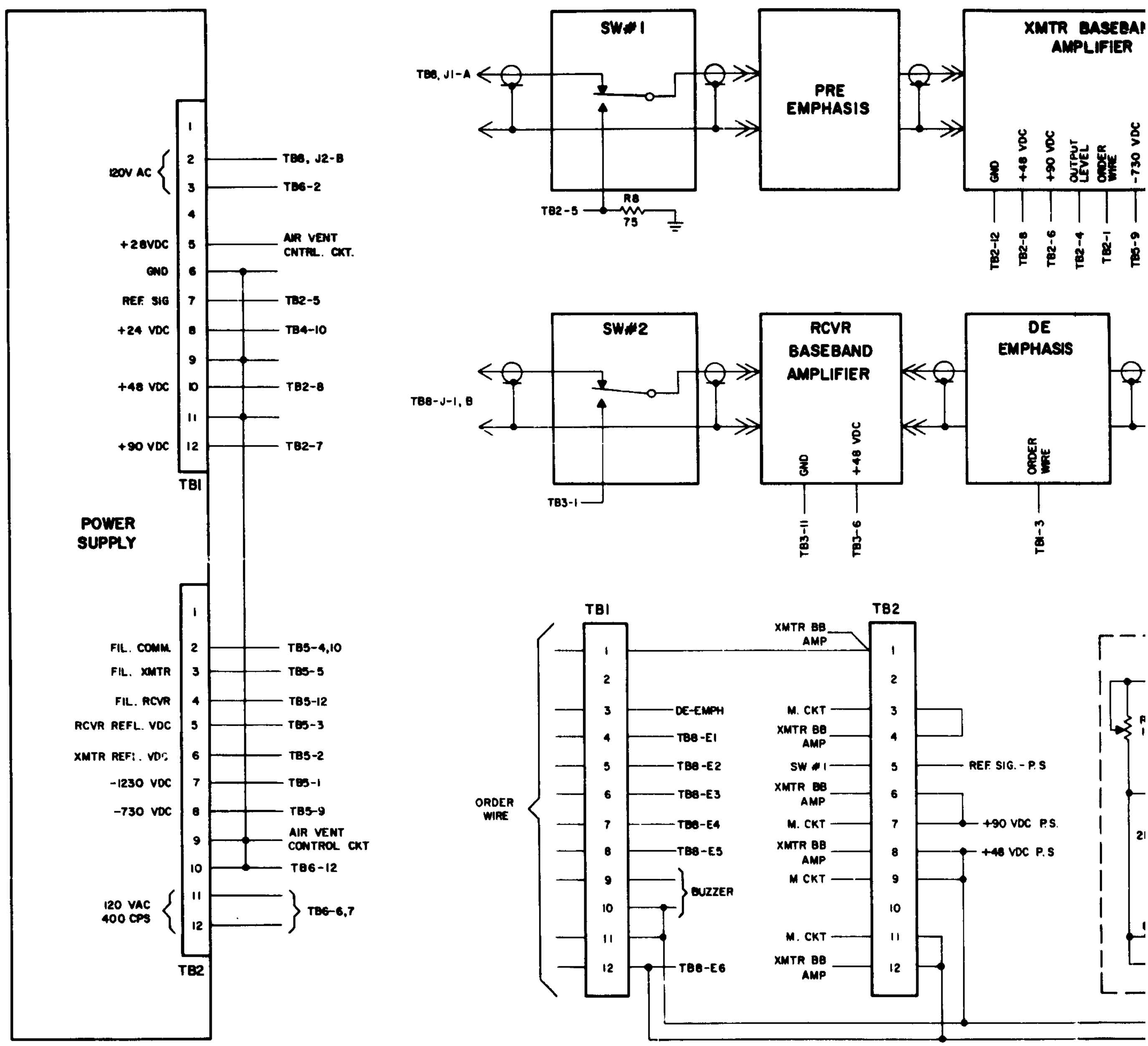


Figure 104 System Interconnection in Laboratory


POWER SUPPLY
TB1
120V AC
+28VDC
GND
REF SIG
+24 VDC
+48 VDC
+90 VDC
TB8, J2-B
TB6-2
AIR VENT CNTRL. CKT.
TB2-5
TB4-10
TB2-8
TB2-7
TB2
FIL. COMM.
FIL. XMTR
FIL. RCVR
RCVR REFL. VDC
XMTR REFL. VDC
-1230 VDC
-730 VDC
120 VAC 400 CPS
TB5-4,10
TB5-5
TB5-12
TB5-3
TB5-2
TB5-1
TB5-9
AIR VENT CONTROL CKT
TB6-12
TB6-6,7
SW#1
TB8, J1-A
TB2-5
R8
75
PRE EMPHASIS
XMTR BASEBAN
AMPLIFIER
GND
+48 VDC
+90 VDC
OUTPUT LEVEL
ORDER WIRE
-730 VDC
TB2-12
TB2-8
TB2-6
TB2-4
TB2-1
TB5-9
SW#2
TB8-J-1, B
TB3-1
RCVR BASEBAND AMPLIFIER
TB3-11
TB3-6
DE EMPHASIS
TB1-3
ORDER WIRE
DE-EMPH
TB8-E1
TB8-E2
TB8-E3
TB8-E4
TB8-E5
BUZZER
TB8-E6
XMTR BB AMP
M. CKT
SW #1
REF SIG. - P.S
+90 VDC P.S.
+48 VDC P.S

1

SHEET I

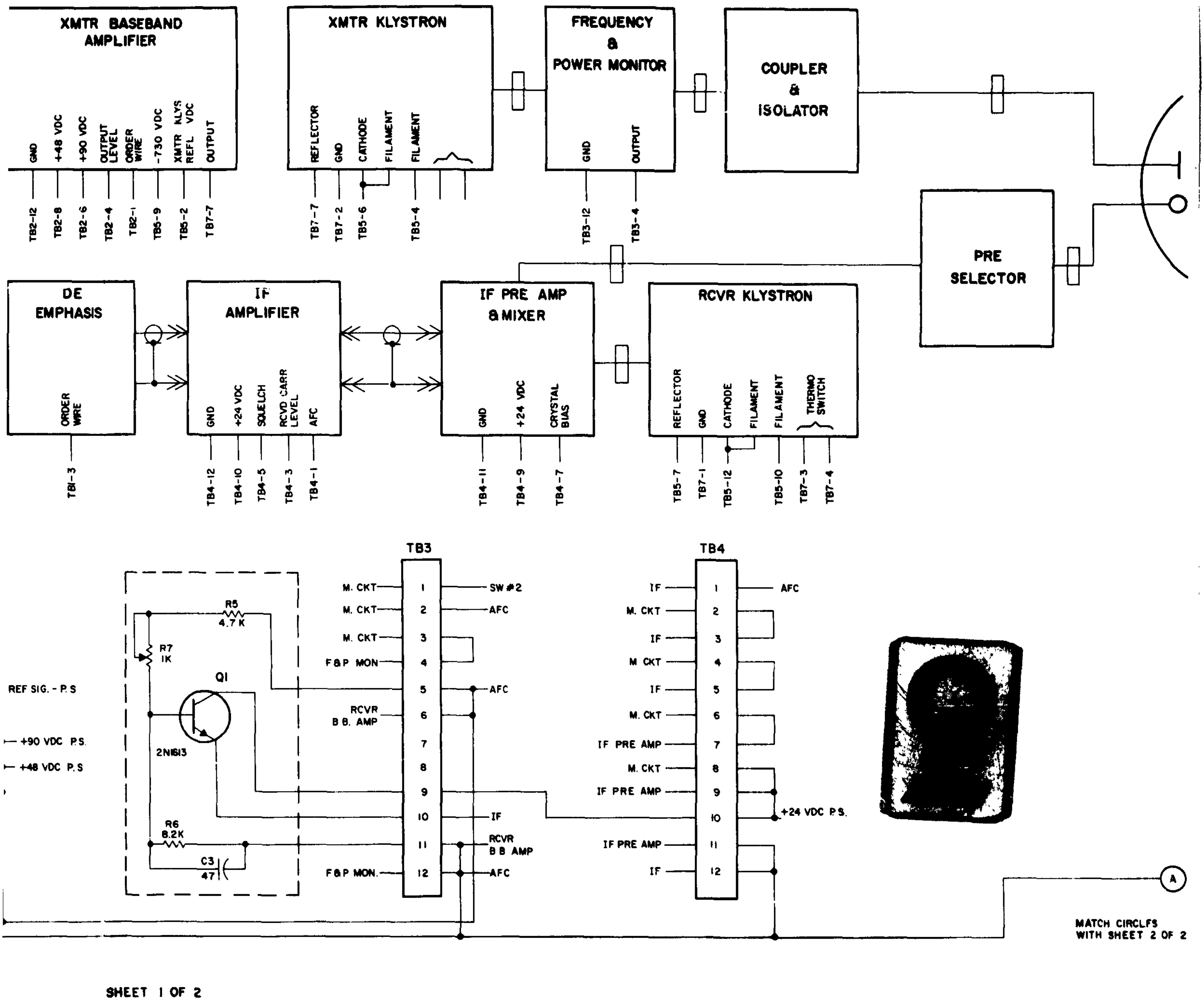
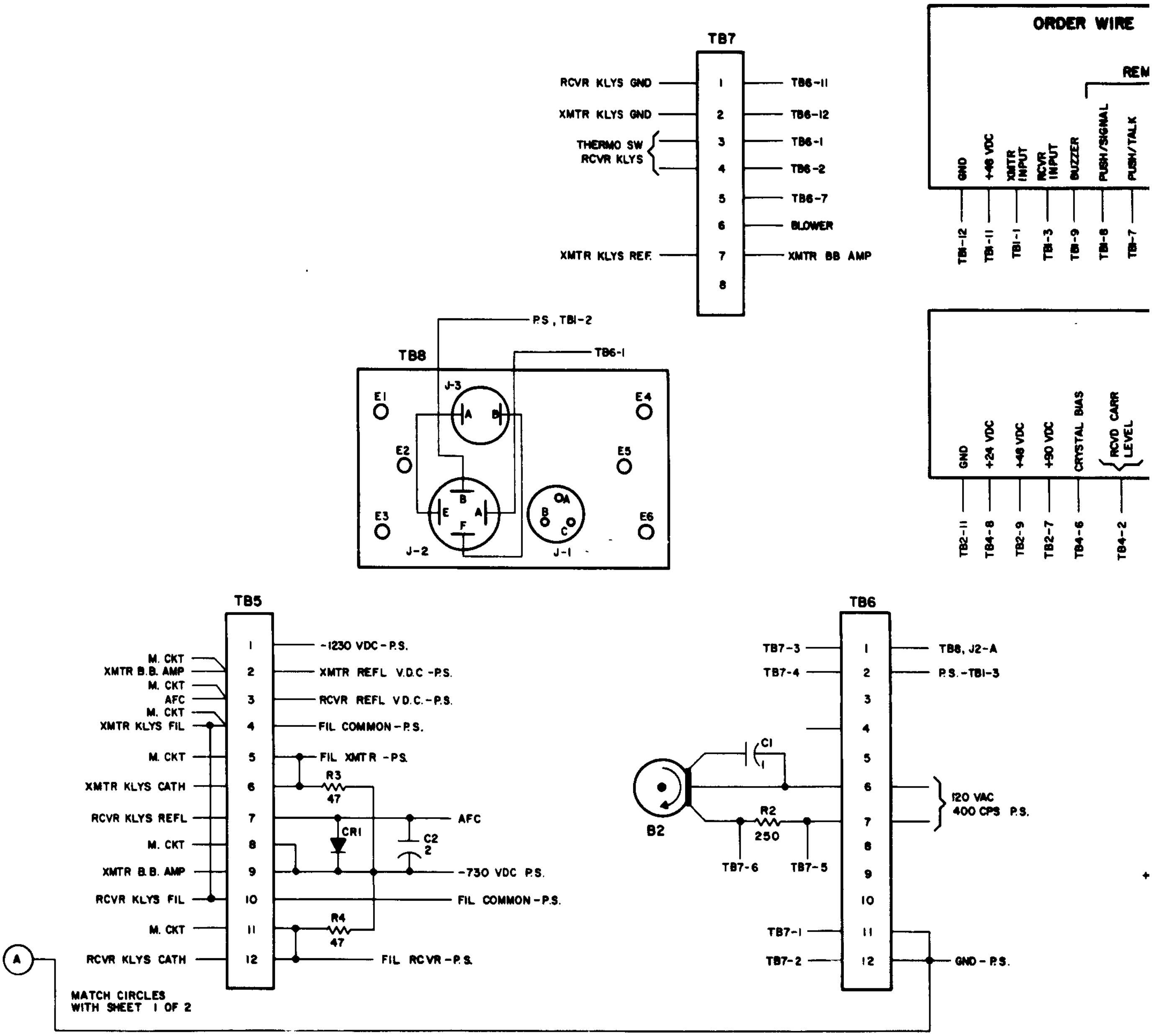
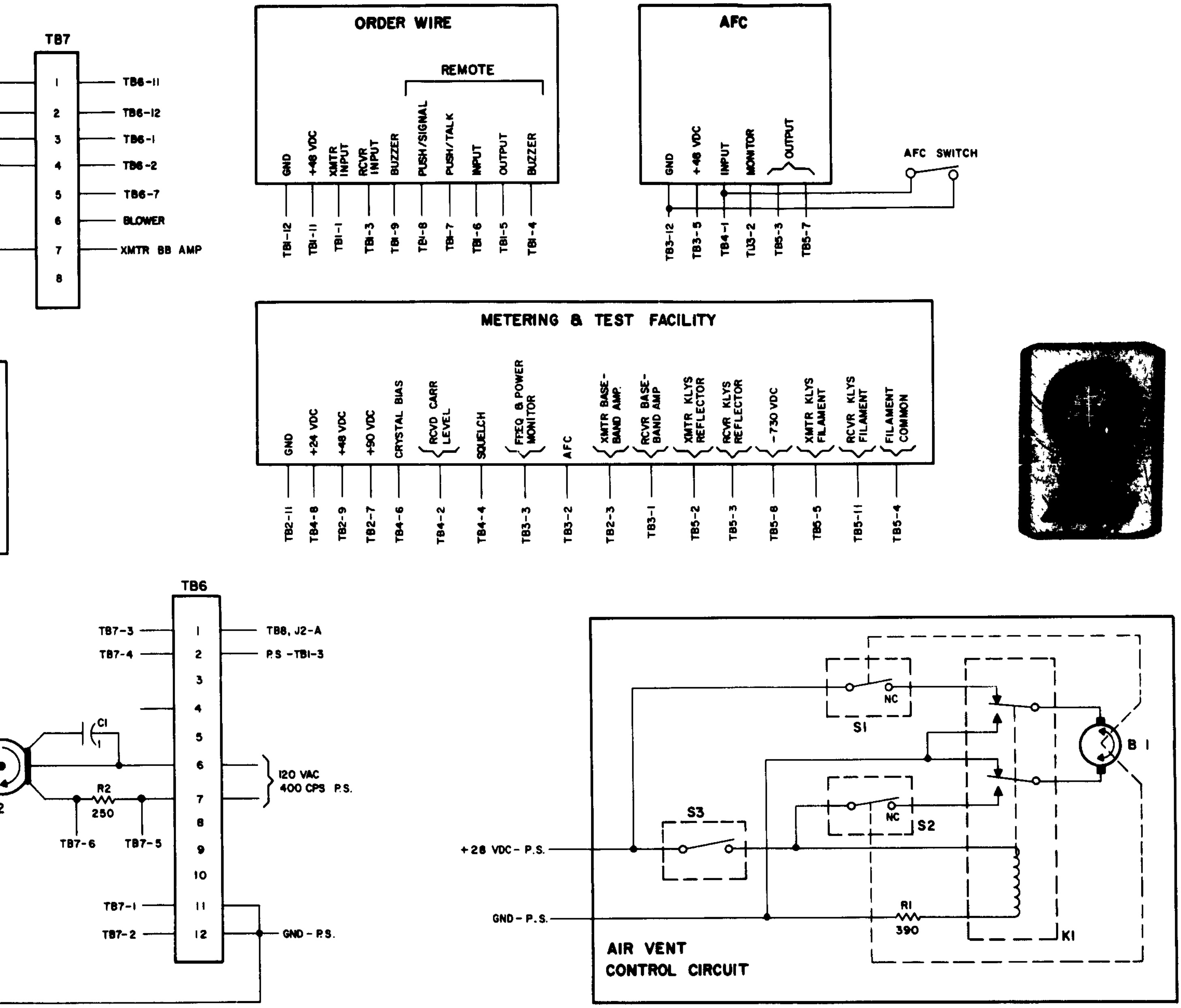


Figure 105 Radio Interconnection Diagram


ORDER WIRE
REM
GND
+48 VDC
XMTR INPUT
RCVR INPUT
BUZZER
PUSH/SIGNAL
PUSH/TALK
TBI-12
TBI-11
TBI-1
TBI-3
TBI-9
TBI-8
TBI-7
GND
+24 VDC
+48 VDC
+90 VDC
CRYSTAL BIAS
RCVD CARR LEVEL
TB2-11
TB4-8
TB2-9
TB2-7
TB4-6
TB4-2
TB7
RCVR KLYS GND
XMTR KLYS GND
THERMO SW
RCVR KLYS
XMTR KLYS REF.
TB6-11
TB6-12
TB6-1
TB6-2
TB6-7
BLOWER
XMTR BB AMP
P.S., TBI-2
TB6-1
TB8
E1
E2
E3
E4
E5
E6
J-3
J-2
J-1
TB5
M. CKT
XMTR B.B. AMP
AFC
XMTR KLYS FIL
XMTR KLYS CATH
RCVR KLYS REFL
RCVR KLYS FIL
RCVR KLYS CATH
-1230 VDC-P.S.
XMTR REFL V.D.C -P.S.
RCVR REFL V.D.C.-P.S.
FIL COMMON-P.S.
FIL XMTR -P.S.
R3
47
CR1
C2
2
-730 VDC P.S.
R4
47
FIL RCVR-P.S.
MATCH CIRCLES WITH SHEET 1 OF 2
TB6
TB7-3
TB7-4
TB8, J2-A
P.S.-TBI-3
C1
B2
R2
250
TB7-6
TB7-5
120 VAC 400 CPS P.S.
TB7-1
TB7-2
GND-P.S.

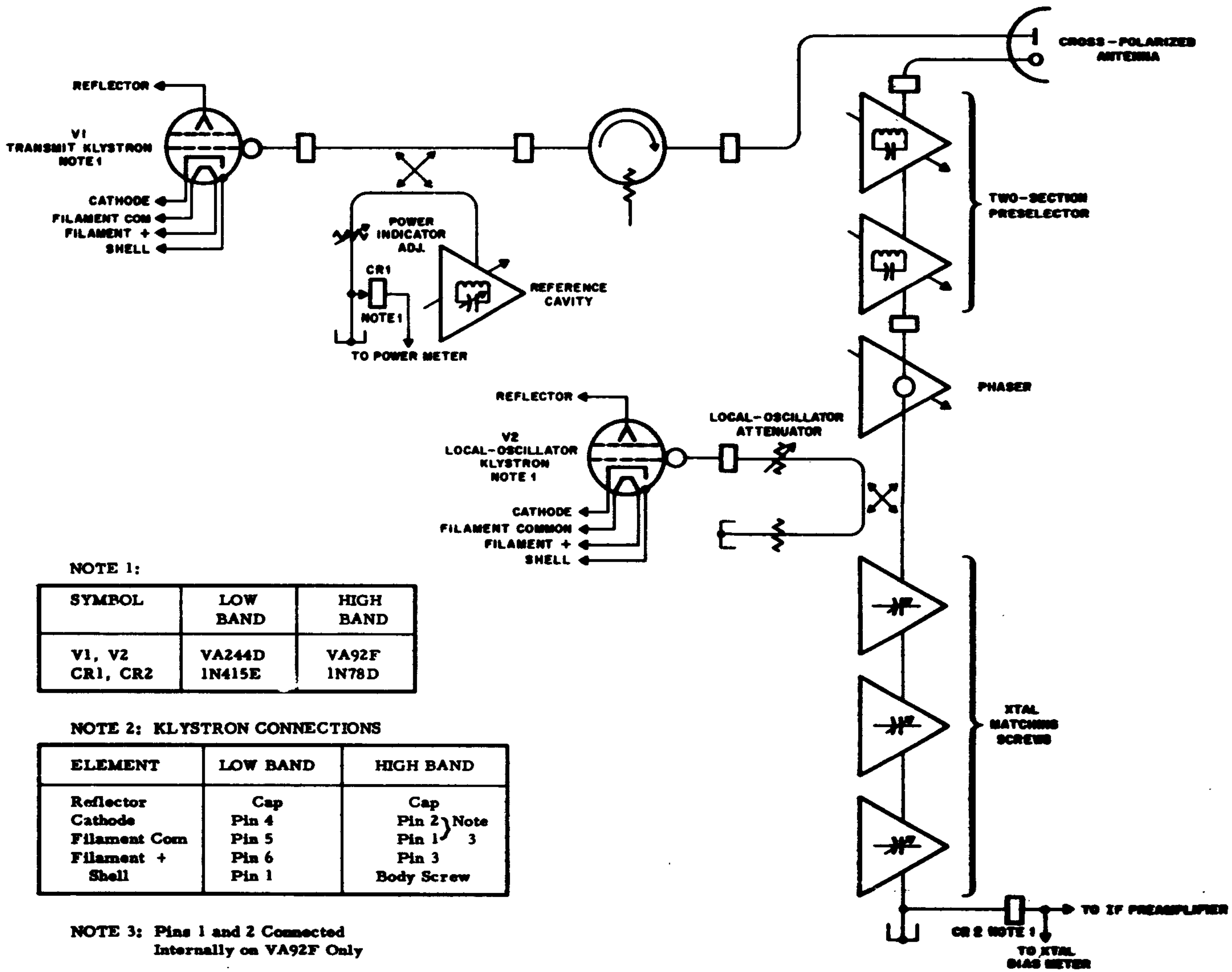
SHEET 2 OF 2

1



SHEET 2 OF 2

Figure 105 Radio Interconnection Diagram



NOTE 1:

SYMBOL	LOW BAND	HIGH BAND
V1, V2 CR1, CR2	VA244D 1N415E	VA92F 1N78D

NOTE 2: KLYSTRON CONNECTIONS

ELEMENT	LOW BAND	HIGH BAND
Reflector	Cap	Cap
Cathode	Pin 4	Pin 2 } Note 3
Filament Com	Pin 5	Pin 1 } Note 3
Filament +	Pin 6	Pin 3
Shell	Pin 1	Body Screw

NOTE 3: Pins 1 and 2 Connected Internally on VA92F Only

Figure 106 AN/TRC-56 Waveguide Schematic

SECTION VII

BIBLIOGRAPHY

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3. CCIR VIII Plenary Assembly, Warsaw; Volume I, 1956.

4. Franke, H.C., "Noise Measurements on Telephone Circuits"; Tele-Tech and Electronic Industries, March 1955.

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6. "IRE Standards on Television: Measurement of Differential Gain and Differential Phase, 1960"; Proceedings of IRE, Volume 48, No. 2, February 1960, pp. 201-208.

7. "Operation and Application of Hybrid Junctions"; The Lenkurt Demodulator, Lenkurt Electronic Company, Incorporated, Volume 2, No. 9, September 1953.

8. Ragazzini, J.R., "The Effect of Fluctuation Voltages on the Linear Detector"; Proceedings of IRE, Volume 30, No. 6, June 1942, pp. 277.

9. "Reference Data for Radio Engineers"; International Telephone and Telegraph Corporation, Fourth Edition, 1956.

10. Rice, S.O., "Mathematical Analysis of Noise"; Bell System Technical Journal, Volume 24, Sections 4.5 and 4.10, January 1945, pp. 46-156.

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11. Sickak, W. and H. Augenblick, "Tuneable Waveguide Filters"; Proceedings of IRE, Volume 39, No. 9, September 1951, pp. 1055-1059.

12. Southworth, G.C., Principles and Applications of Waveguide Transmission; D. Van Nostrand Company, Incorporated, 1953.

APPENDIX I

FREQUENCY DEVIATION CALCULATIONS FOR AN/TRC-56
(By Q. Veit and R. Smith)

A. SPECIFICATIONS AND EQUIPMENT CHARACTERISTICS

1. General

Two important factors affect the frequency deviation required in the RF unit of the AN/TRC-56. These are the overall systems and equipment specifications as outlined in Exhibit RADC-2562A and the equipment design approach chosen within the confines of Exhibit 2562A. The pertinent system and equipment specifications and characteristics dictated by Exhibit 2562A and the design approach are detailed in the following sections.

2. Exhibit RADC 2562A Specifications

Exhibit RADC 2562A states:

"3.5.2.8.2 - RECEIVER NOISE FIGURE - The receiver (RF) noise figure shall be less than 10 db."

"3.5.2.8.4 - INTERMEDIATE-FREQUENCY (I-F) AND DISCRIMINATOR BANDWIDTH - The i-f discriminator bandwidth shall be 20 MC to the minus 3 db points, less than 30 MC to the minus 20 db points, and less than 50 MC to the minus 60 db points."

"3.5.2.10.6 - CARRIER-TO-NOISE RATIO - The carrier-to-noise ratio (I.F.) of the system shall not be less than 40 db with a 60 db path loss inserted."

"3.5.3.1 - DATA CHANNEL MULTIPLEX - The data channel multiplex shall provide facilities for simultaneous transmission and reception of twelve 20 KC channels in each direction over the microwave relay system."

"3.5.3.2 - TWELVE-CHANNEL VOICE MULTIPLEX - The 12 channel voice multiplex shall provide facilities for transmission and reception of 12 voice channels over the microwave in each direction."

"3.5.3.2.1 - AMPLITUDE VERSUS FREQUENCY RESPONSE - Each channel of the voice multiplex shall be capable of transmitting frequencies between 300 and 3500 cps . . ."

"3.5.3.2.12 - SYSTEM NOISE - The noise measured at the receiving end, when converted to the 0 db relative transmission level point, shall not exceed 32 dbs (decibels adjusted) as measured with Transmission Measuring Set TS-559/FT (or equal) using FIA weighting."

3. Pertinent Equipment Characteristics Resulting from Design Approach

The following multiplex equipment characteristics are a result of the design approach:

Modulation	-	Double sideband transmitted carrier.
Modulation Index	-	Thirty percent for standard channel test tone.
Carrier Frequencies	-	50 to 1200 kc inclusive, spaced at 50-kc intervals.

B. REQUIRED MULTIPLEX RECEIVER CHANNEL INPUT S/N RATIO

1. Conversion of FIA Weighted Noise to Flat S/N Ratio

Exhibit RADC 2562A requires that the multiplex channel output noise be 32 dba (FIA weighting) relative to the zero db relative transmission level point. At the zero transmission level point, the standard channel test tone is 0 dbm or 85 dba. This gives:

$$S/N = 85 \text{ dba} - 32 \text{ dba} = 53 \text{ db (FIA weighting)} \quad (1)$$

Converting the FIA weighted signal-to-noise ratio into a flat-weighted signal-to-noise ratio gives:[2]

$$S/N \text{ Flat} = S/N \text{ FIA} - 3 \text{ db} = 53 - 3 = 50 \text{ db} \quad (2)$$

NOTE: The above numbers are based on measuring the noise in a 3-kc audio bandwidth from 300 cps to 3300 cps.

2. Effect of Multiplex Receiver Demodulation Process

The detector which removes the amplitude modulation from the subcarrier in each channel is assumed to be a diode detector. The channel output signal-to-noise ratio is given by the following formular (provided the signal-to-noise ratio in the input of the detector is large): [3,4]

$$R_o = \frac{kR}{1 + k^2/2} \tag{3}$$

where:

R_o = Output signal-to-noise voltage ratio

R = Input (unmodulated carrier) signal-to-noise voltage ratio

k = Modulation index

For 30 percent modulation:

$$R_o = 0.3\ R$$

and the decrease in signal-to-noise ratio resulting from the detection process is approximately 10 db. Therefore, to provide a multiplex audio output signal-to-noise ratio of 50 db, the unmodulated carrier-to-noise ratio at the input of the multiplex channel receiver must be 60 db.

NOTE: The input signal-to-noise ratio R is defined as the ratio of the unmodulated carrier voltage at the multiplex receiver input to the noise voltage in the baseband spectrum occupied by the multiplex channel.

C. THEORETICAL DEVIATION REQUIREMENTS

1. General

The carrier-(RF)-to-noise ratio specified in paragraph 3.5.2.10.6 of exhibit RADC 2562A is a function of the received RF carrier power, the IF bandwidth, and the receiver noise figure. For the purpose of calculating the required channel deviation, it is assumed that the specified carrier-to-noise ratio of 40 db represents the conditions existing during median RF

propagation periods. It is also assumed that under these conditions the multiplex channel signal-to-noise ratio of 50 db as determined in Section B-1 must be met.

2. Single Channel Deviation

The unmodulated channel subcarrier-to-noise ratio at the multiplex detector input is given by:[4]

$$\frac{E_s}{E_n} = \sqrt{\frac{S}{2KTB_cW}} \cdot \frac{\Delta F}{f_c} \qquad (4)$$

where:

S = Input RF carrier power in watts

KT = 4×10^{-21} watts/cps

W = Microwave receiver noise figure

B_c = Multiplex channel bandwidth

f_c = Multiplex subcarrier frequency

ΔF = Peak deviation of RF carrier due to the unmodulated subcarrier

Equation (4) may be written as:

$$\frac{E_s}{E_n} = \sqrt{\frac{P_{cn} \quad B_i}{2B_c}} \cdot \frac{\Delta F}{f_c} \qquad (5)$$

where:

P_{cn} = RF carrier-to-noise power ratio in IF

B_i = Bandwidth of the microwave IF

B_c, ΔF, f_c = As defined before.

Using:

$B_c = 6$ kc

$B_i = 20$ Mc

$P_{cn} = 10^4$ (40 db)

$E_s/E_n = 10^3$ (60 db)

and solving Equation 5 for ΔF gives:

$$\Delta F = 0.245 \quad f_c \tag{6}$$

Equation 6 is plotted as Curve No. 2 in Figure I-1.

3. Multiple Channel Deviation

The peak deviation of the RF carrier by a single subcarrier is given by Equation 6. The rms RF deviation due to a single subcarrier is:

$$\Delta F \text{ rms} = \frac{0.245 \quad 2 \quad f_c}{2} \tag{7}$$

Using:

$$f_c = 50 \times 10^3 n \text{ cps} \tag{8}$$

where: n = Channel number (1 to 24)

Equation 7 becomes:

$$\Delta F \text{ rms} = 8.7 \times 10^3 n \text{ cps} \tag{9}$$

The total rms RF deviation due to all 24 channel subcarriers is given by:

$$\text{Total } \Delta F \text{ rms} = 8.7 \times 10^3 n \sqrt{\sum_1^{24} n^2} \text{ cps} \tag{10}$$

$$= 609 \text{ kc} \tag{10a}$$

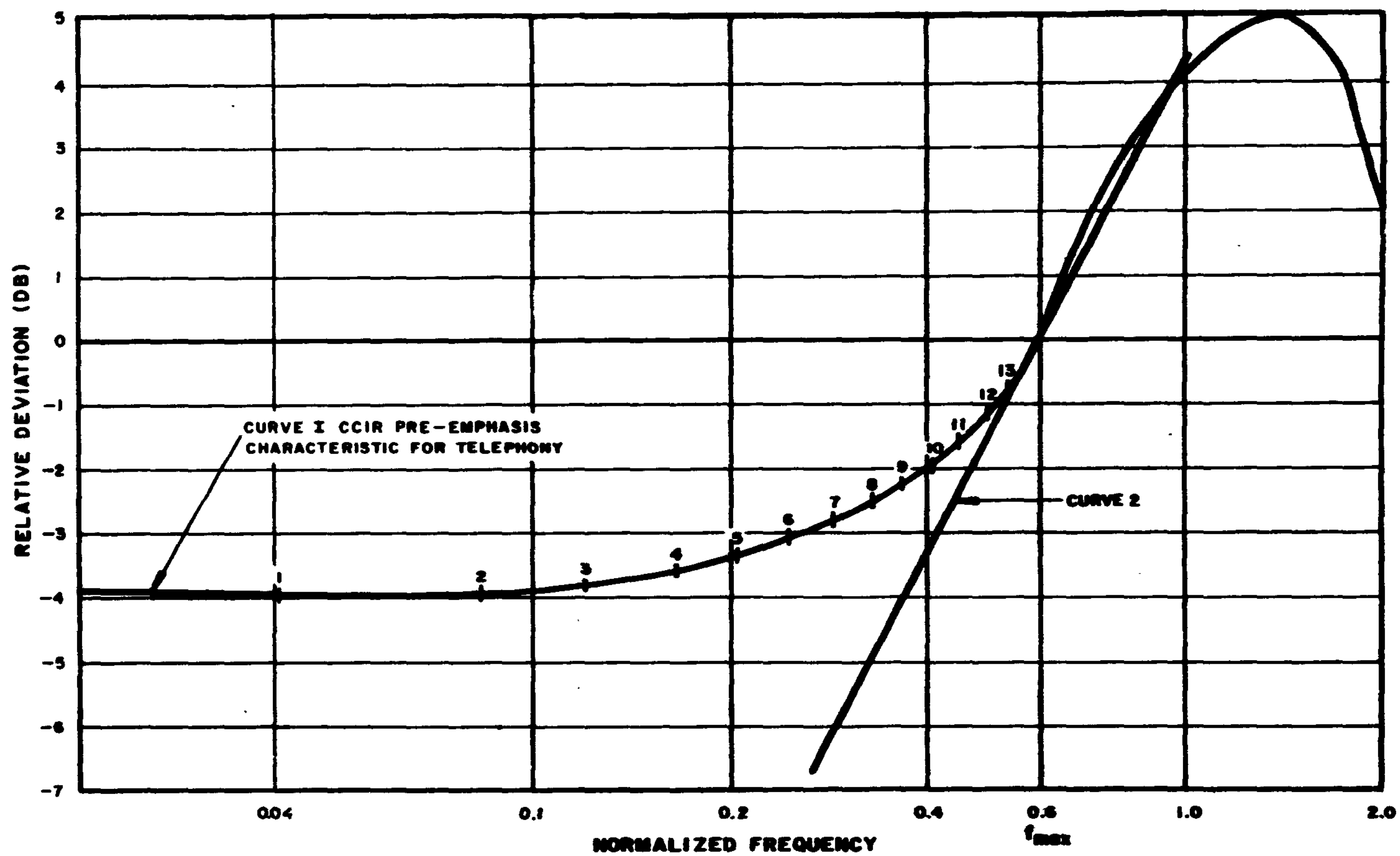

Figure I-1 Pre-emphasis Curves

Combining Equations 6 and 8:

$$\Delta F_n = 12.25 \times 10^3 n \text{ cps} \tag{11}$$

where:

ΔF_n = Peak RF carrier deviation due to the unmodulated subcarrier of the n^{th} multiplex channel

n = Multiplex channel number (1 to 24)

Letting K be the RF transmitter modulation sensitivity (cps/volt):

$$E_{pn} = \frac{\Delta F_n}{K} = \frac{12.25 \times 10^3 n}{K} \text{ volts} \tag{12}$$

where:

E_{pn} = Peak voltage at the RF transmitter input due to the unmodulated subcarrier on the n^{th} multiplex channel

K = RF transmitter modulation sensitivity (cps/volt)

The instantaneous voltage expression for a single unmodulated subcarrier at the RF transmitter input may be written as:

$$E_n = E_{pn} \cos (n\omega t + \phi n) \tag{13}$$

where:

E_n = Instantaneous voltage of n^{th} multiplex channel unmodulated subcarrier

E_{pn}, n = As defined before

ω = 2π (50×10^3) radians/second

ϕn = An arbitrary phase angle

If all the channel subcarriers are generated from a common source with the same time delay for all carriers, then ϕn may be set equal

to zero without any loss of generality. Using this assumption, the expression for the composite signal (24 subcarriers) at the RF transmitter input becomes:

$$E_{p+} = \sum_{n=1}^{24} E_{pn} \cos(n\omega t) \tag{14}$$

The expression in Equation 14 will have a positive maximum where cos ($n\omega t$) equals plus one for all values of n from 1 to 24. This will occur at $t = 2k\pi/w$; $k = 1, 2, 3, \ldots$. At these points the composite voltage will have a peak of:

$$E_{p+} = \sum_{n=1}^{24} E_{pn} = \frac{12.25 \times 10^3}{K} \times \sum_{n=1}^{24} n \text{ volts} \tag{15}$$

Multiplying the positive peak voltage by K to obtain the positive peak frequency deviation gives:

$$\text{Peak } \Delta F_+ = KE_{p+} = 3.68 \times 10^6 \text{ cps} \tag{16}$$

D. PRACTICAL DEVIATIONS

The deviations calculated in Sections C-2 and C-3 are based on the assumption that the channel signal-to-noise degradation is due only to intrinsic (idle) microwave receiver noise. In a practical system, this assumption is valid for the channels occupying the upper half of the baseband spectrum used by the multiplex. For the lower half of the channels, additional channel signal-to-noise degradation is caused by residual FM noise in the microwave transmitter, receiver imperfections, and amplitude nonlinearities (crosstalk) in the equipment.

This subject has been thoroughly explored by CCIR[4] and the CCIR-recommended pre-emphasis curve shown in Figure I-1 has resulted. Plotted along with the CCIR pre-emphasis curve is the theoretical pre-emphasis curve arrived at from the calculations in Sections C-2 and C-3. CCIR recommends that the multiplex channel falling at 0.608 f_{max} (f_{max} - highest channel frequency) have the same deviation with and without pre-emphasis. Observation of the two curves shows the CCIR and theoretical curves to coincide for frequencies above 0.608 f_{max}.

Applying the CCIR pre-emphasis curve to the particular case being considered here results in:

$$\Delta F_n = 12.25 \times 10^3 \text{ n cps} \tag{17}$$

for n = 14 to n = 24

The peak deviations for the channels 1 to 13 become:

Channel	Frequency (kc)	Peak Deviation (kc)
1	50	108
2	100	110
3	150	112
4	200	113
5	250	117
6	300	120
7	350	124
8	400	129
9	450	133
10	500	137
11	550	144
12	600	151
13	650	160
Total peak deviation for channels 1 to 13 =		1658

The total peak deviation for channels 14 to 24 is:

$$F_{(14\text{-}24)} = \sum_{n=14}^{24} 12.25 \times 10^3 n = 2560 \text{ kc} \tag{18}$$

The total peak deviation for all 24 channels then is:

$$F_T = 2560 \text{ kc} + 1658 \text{ kc} \tag{19}$$

$$= 4.218 \text{ Mc}$$

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APPENDIX II

SIGNAL-TO-NOISE CALCULATIONS FOR AN/TRC-56
(By Q. Veit and R. Tamoshunas)

A. GENERAL

In any radio system, two important factors to be considered when computing the system's overall signal-to-noise ratio are: thermal noise and intermodulation or crosstalk noise. The transmit baseband amplifier, transmit klystron, receiver IF (includes RF mixer), IF baseband amplifier, and receive baseband amplifier are the units in AN/TRC-56 Radio System where the contribution of the signal-to-noise ratio occurs (see Figure II-1). Effects of thermal and intermodulation noise at each unit will be discussed, and overall signal-to-noise ratio of the system computed in the following sections of this appendix.

B. SPECIFICATIONS AND EQUIPMENT

1. Specifications

Exhibit RADC-2562A states:

"3.5.2.8.2 - RECEIVER NOISE FIGURE - The receiver (RF) noise figure shall be less than 10 db."

"3.5.2.10.6 - CARRIER-TO-NOISE RATIO - The carrier-to-noise ratio (I.F.) of the system shall not be less than 40 db with a 60 db path loss inserted."

"3.5.3.1.6 - CROSS-TALK - The cross-talk coupling between the 20 KC channels on a single link shall be numerically less than 60 db, when referred to equal level points."

"3.5.3.2.12 - SYSTEMS NOISE - The noise measured at the receiving end, when converted to the 0 db relative transmission level point, shall not exceed 32 dba (decibels adjusted) as measured with Transmission Measuring Set TS-559/FT (or equal) using FIA weighting."

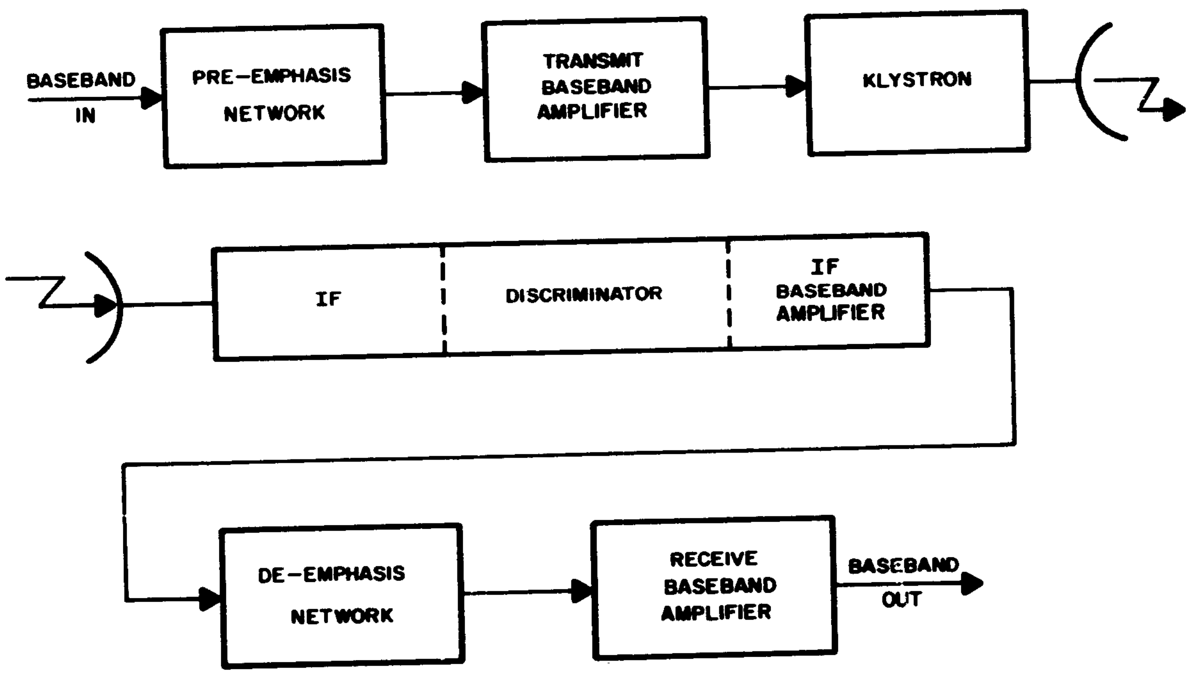

Figure II-1 Block Diagram of AN/TRC-56 Radio System

2. Proposed Equipment Characteristics

The following are the proposed equipment characteristics:

a. Klystron modulation sensitivity, 400 kc/volt.

b. Transmit baseband amplifier voltage gain of 30 db.

c. Transmit and receive baseband amplifier input impedance of 75 ohms.

d. IF baseband amplifier input impedance of 2000 ohms.

e. IF baseband amplifier voltage gain of 12 db.

f. B_c - multiplex channel bandwidth of 6 kc.

g. Pre-emphasis network with a pivot point at 700 kc frequency channel (CCIR Recommendation No. 275).

h. Twenty-four multiplex channels, from 50 kc to 1200 kc in 50-kc intervals.

C. THERMAL NOISE

1. General

The total signal-to-noise ratio in the system due to thermal agitation can be expressed as:

$$S/N_T = \frac{1}{N_{tbt}/S + N_{kt}/S + N_{ift}/S + N_{ibt}/S + N_{rbt}/S} \tag{1}$$

where:

N_{tbt} = Thermal noise in transmit baseband amplifier

N_{kt} = Thermal noise in transmit klystron

N_{ift} = Thermal noise in IF amplifier

N_{ibt} = Thermal noise in IF baseband amplifier

N_{rbt} = Thermal noise in receive baseband amplifier

S = Signal power at noise measuring points

The thermal noise in the transmit baseband amplifier, klystron and IF baseband amplifier is independent of frequency. The baseband signal in the above-mentioned units is pre-emphasized and, therefore, S/N_T degradation in these units will be worst at the lowest frequency channel. The S/N_T in the receive baseband amplifier will be the same for all channels, and in the IF will be worst at the highest frequency channel.

2. Thermal Noise Levels at Each Unit

a. Transmit Baseband Amplifier

A multiplex channel S/N_T of 50 db is required to meet RADC 2562A, 3.5.3.2.12. The required deviation to meet a 50 db S/N_T ratio has been calculated, assuming only the thermal noise of the IF to be significant (see Appendix I).

The deviation is:

$$\Delta F = 0.245 \ f_c, \text{ for channels above } 0.6 \ f_c \text{ maximum} \qquad (2)$$

where:

ΔF = peak deviation of RF carrier due to the unmodulated multiplex subcarrier

f_c = subcarrier frequency

The deviation of the lowest channel will be 4 db (0.63) lower than that of a channel at 0.6 f_c maximum (CCIR Recommendation No. 275),

or:

$$\Delta F_1 = (0.245)\ (0.6)\ (1200 \text{ kc})\ (0.63) = 111 \text{ kc} \qquad (3)$$

Since the klystron modulation sensitivity is 400 kc/volt, the signal at its reflector is:

$$E_r = \frac{111 \text{ kc}}{400 \text{ kc/volt}} = 0.278 \text{ volt peak} \tag{4}$$

For a baseband amplifier gain of 30 db (32:1), the channel carrier input voltage is:

$$E_{in} = \frac{0.278 \text{ volt}}{32} = 0.00868 \text{ volt peak, or } 0.00615 \text{ volt rms} \tag{5}$$

From this, the input signal power is found to be:

$$P_{in} = \frac{(E_{in})^2}{R} \text{ watts}$$

$$= 10 \log \frac{(6.15 \times 10^{-3})^2}{75} \text{ dbw} = -63 \text{ dbw} \tag{6}$$

Assume that the transmit baseband amplifier has a 10-db noise figure. Then the effective noise power in a 6 kc bandwidth at the baseband amplifier input is:

$$N_{tbt} = \left[N_{in} + 10 \log B_c + W_n \right] \text{ dbw} \tag{7}$$

where:

N_{in} = Thermal noise power/cps (-204 dbw)

B_c = Multiplex channel bandwidth (cps)

W_n = Noise figure of the baseband amplifier (db)

Then:

$$N_{tbt} = \left[-204 + 10 \log (6 \times 10^3) + 10 \right] \text{ dbw} \tag{8}$$

$$= -156 \text{ dbw}$$

From Equations 6 and 8, the effective signal-to-noise ratio of the transmit baseband amplifier is:

$$S/N_{tbt} = -63 \text{ dbw} - (-156 \text{ dbw}) = 93 \text{ db} \tag{9}$$

<u>Conclusions:</u> Thermal noise in the transmit baseband amplifier may be ignored, since the signal-to-noise ratio due to it alone is 33 db better than required. If modulation sensitivity of the klystron increases, the 33-db margin increases in direct proportion.

b. <u>Transmit Klystron</u>

The thermal amplitude modulation noise in the klystron is 60 db or more below the carrier so that its effect on the carrier-to-noise ratio in the IF is negligible. Based on past experience, the residual frequency modulation noise in the klystron is also assumed to be negligible.

c. <u>IF Baseband Amplifier</u>

Thermal noise in the IF baseband amplifier will have the worst effect at the lowest frequency channel. The deviation for this channel carrier (Equation 3) is 111 kc peak.

The signal voltage out of the discriminator will be:

$$E_o = \frac{111 \text{ kc}}{\sqrt{2}} \times S_D \text{ volts rms} \tag{10}$$

where:

S_D = Discriminator slope in volts/kc

Signal power at this point will be:

$$P_o = \frac{(E_o)^2}{R} = \frac{(111 \text{ kc} \times S_D)^2}{2R} \text{ watts} \tag{11}$$

Assuming a 10-db noise figure for the IF baseband amplifier, the noise power in the lowest frequency channel will be -156 dbw, from Equation 8.

The noise power can be expressed as:

$$N_{ift} = \frac{(E_n \text{ rms})^2}{R} \text{ in watts}$$

or,

$$N_{ift} = 10 \log \frac{(E_n \text{ rms})^2}{R} \text{ in dbw} \tag{12}$$

$$N_{ift} = 10 \log \frac{(E_n \text{ rms})^2}{R} = -156 \text{ dbw} \tag{13}$$

$$(E_n \text{ rms})^2 = 3.98 \times 10^{-16} R \tag{14}$$

Assuming that the IF baseband amplifier input impedance is 2000 ohms and that the required effective signal-to-noise ratio in the amplifier is higher than 80 db or 10^8:1 ratio, the signal-to-noise voltage ratio can be expressed as:

$$\frac{E_s \text{ rms}}{E_n \text{ rms}} = \frac{111 \text{ kc} \times S_D/\sqrt{2}}{\sqrt{3.98 \times 10^{-16} R}} = \sqrt{10^8} \tag{15}$$

then:

$$S_D = \frac{10^4 \times \sqrt{3.98 \times 10^{-16} \times 2 \times 10^3} \times \sqrt{2}}{111}$$

$$= 0.113 \times 10^{-3} \text{ volt/kc} \tag{16}$$

or,

0.113 volt/Mc

If the discriminator slope is 0.113 volt/Mc, then the thermal noise of the IF baseband amplifier is over 20 db below the thermal noise out of the IF. If this slope is decreased to 30 mv/Mc, then the margin has decreased to approximately 10 db and the overall system signal-to-noise will be degraded less than 0.4 db. If the IF baseband amplifier input impedance is increased from the assumed value of 2000 ohms by a factor of 2, the discriminator slope must be increased by a factor of $\sqrt{2}$ to maintain the same signal-to-noise ratio.

<u>Conclusion:</u> Ignore thermal noise of the IF video amplifier with the above reservations.

d. <u>Receive Amplifier</u>

Assuming a 10-db noise figure for the receive baseband amplifier, the effective noise power at the input in the lowest frequency channel will be -156 dbw, from Equation 8. The lowest frequency channel is attenuated 1.0 db by the de-emphasis network and amplified 12 db by the IF baseband amplifier. If the discriminator slope (S_D) is 30 mv/Mc, then the rms voltage of the lowest frequency channel carrier at the receive baseband amplifier input is:

$$E_{rms} = \left[\frac{(\Delta F_1)}{\sqrt{2}} (S_D) \right] (4)\ (0.9)$$

$$= \frac{(111\text{ kc})\ (30 \times 10^{-3})\ (4)\ (0.9)}{\sqrt{2}\qquad (1000\text{ kc})}$$

$$= 8.43 \times 10^{-3}\text{ volts} \tag{17}$$

Signal power is:

$$P_s = 10 \log \left[\frac{(E_{rms})^2}{R} \right]$$

$$= 20 \log (8.43 \times 10^{-3}) - 10 \log 75 \tag{18}$$

$$P_s = -60.5\text{ dbw} \tag{19}$$

The signal-to-noise ratio in the receive baseband amplifier is:

$$S/N_{rbt} = -60.5 - (-156) = 95.5\text{ db} \tag{20}$$

<u>Conclusion</u>: Ignore thermal noise in the receive baseband amplifier. Even if the IF baseband amplifier voltage gain is unity and it is used only as an impedance transformer, the S/N will be more than 83.5 db.

3. <u>Conclusion</u>

With the previous assumption that only the thermal noise of the IF is significant (calculations for the IF thermal noise have been made

in Appendix I), it was proven that the signal-to-thermal noise ratio in the transmit baseband amplifier, klystron, IF baseband amplifier, and receive baseband amplifier are at least 10 db higher than that of the IF. Therefore, the only thermal noise that must be considered is that of the IF amplifier.

D. INTERMODULATION OR CROSS-TALK NOISE

1. General

The total signal-to-noise ratio in the system due to intermodulation can be expressed as:

$$S/N_I = \frac{1}{N_{tbi}/S + N_{ki}/S + N_{ifi}/S + N_{ibi}/S + N_{rbi}/S} \quad (21)$$

where:

N_{tbi} = Cross-talk in transmit baseband amplifier

N_{ki} = Cross-talk in transmit klystron

N_{ifi} = Cross-talk in receive IF

N_{ibi} = Cross-talk in IF baseband amplifier

N_{rbi} = Cross-talk in receive baseband amplifier

S = Signal power at noise measuring points

The intermodulation has been assigned to the components of the system in the following manner:

a. Transmit baseband amplifier	1
b. Transmit klystron	4
c. Total IF	2
(1) IF discriminator	1
(2) IF baseband amplifier	1
d. Receive baseband amplifier	1

The overall channel signal-to-cross-talk ratio is 60 db. This means a channel carrier-to-cross-talk ratio of 70 db (Appendix I). With the above proportions, the channel-to-cross-talk ratios must be:

a.	Transmit baseband amplifier	79 db
b.	Transmit klystron	73 db
c.	Total IF	76 db
	(1) IF discriminator	79 db
	(2) IF baseband ampifier	79 db
d.	Receive baseband amplifier	79 db

2. Intermodulation Noise Computation

To establish conveniently measurable design goals, the signal-to-cross-talk ratio is converted into noise-to-noise ratio. The multiplex used in the AN/TRC-56 system occupies a frequency spectrum from 50 kc to 1200 kc. This spectrum is equivalent to that occupied by a 240-channel single-sideband multiplex, and it is reasonable to use noise-loading techniques to evaluate the intermodulation. The peak random noise-loading signal should be set equal to the peak multiplex signal.

From Appendix I:

$$\Delta F_n = 12.25 \times 10^3 n \text{ cps} \tag{22}$$

where:

ΔF_n = Peak deviation of RF carrier due to the unmodulated subcarrier at the n^{th} channel

n $\geq$ 14, above the 700-kc frequency channel

$$\Delta F_{14} = 12.25 \times 10^3 \times 14 = 171.5 \text{ kc} \tag{23}$$

Also from Appendix I, the total peak deviation is:

$$\Delta F_T = \Delta F_{(1 - 14)} + \Delta F_{(14 - 24)} = 4.218 \text{ Mc peak} \tag{24}$$

The peak voltage required at the klystron reflector is:

$$E_{REF} = \frac{4.218 \text{ Mc}}{0.4 \text{ Mc/volt}} = 10.5 \text{ volts peak} \tag{25}$$

Considering random noise to have a 13-db (4.46:1) peak factor, the following relationship holds:

$$E_n \text{ rms} = \frac{E_{np}}{4.46} \tag{26}$$

Setting the noise test signal peak voltage equal to the peak multiplex signal at the klystron reflector (Equation 25) gives:

$$E_n \text{ rms} = \frac{10.5}{4.46} = 2.35 \text{ volts rms} \tag{27}$$

The pre-emphasis network is such that when a random noise signal is passed through it, the noise power in the frequency range 700 to 1200 kc is increased as much as the noise power in the frequency range 50 to 700 kc is decreased. This makes 700 kc a pivot point (average) such that the total noise output of the pre-emphasis network is given by the noise power in a given channel bandwidth centered at 700 kc multiplied by the number of such channel bandwidths which are contained in the 50- to 1200-kc frequency range. The noise test signal at the klystron reflector in a 6-kc bandwidth centered at 700 kc (channel 14) is given by:

$$N_s = \frac{E_n \text{ rms} \times B_c}{B_t}$$

$$= \frac{2.35 \times 6 \text{ kc}}{(1200\text{-}50) \text{ kc}} = 0.012 \text{ volt rms} \tag{28}$$

where:

B_c = Channel bandwidth (6 kc)

B_t = Total noise test signal bandwidth

The 700-kc channel carrier voltage present at the klystron reflector is (Equation 23):

$$E_c = \frac{\Delta F_{14}}{\sqrt{2} \times 0.4 \text{ Mc/volt}} = 0.303 \text{ volt rms} \qquad (29)$$

The intermodulation noise (cross-talk) voltage in a 6-kc slot measured at the 700-kc channel must be 79 db below the channel carrier rms voltage at the klystron reflector (transmit baseband amplifier output). Therefore,

$$20 \log (E_c/N_I) = 79 \text{ db} \qquad (30)$$

or

$$N_I = 0.303 \times 0.112 \times 10^{-3} = 0.034 \text{ mv rms} \qquad (31)$$

Using Equations 28 and 31, the ratio of the noise test signal in a 6-kc band centered at 700 kc to the intermodulation noise present in the same band (the noise power ratio) is:

$$N_s/N_I = \frac{12 \times 10^{-3}}{0.034 \times 10^{-3}} \qquad (32)$$

This noise power ratio must be obtained for all channels.

3. Conclusions

From the preceding calculations, the noise-to-noise ratio for the following units of AN/TRC-56 system should be:

a.	Transmit baseband amplifier	50 db
b.	Transmit klystron	44 db
c.	Total IF	47 db
	(1) IF discriminator	50 db
	(2) IF baseband amplifier	50 db
d.	Receive baseband amplifier	50 db

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